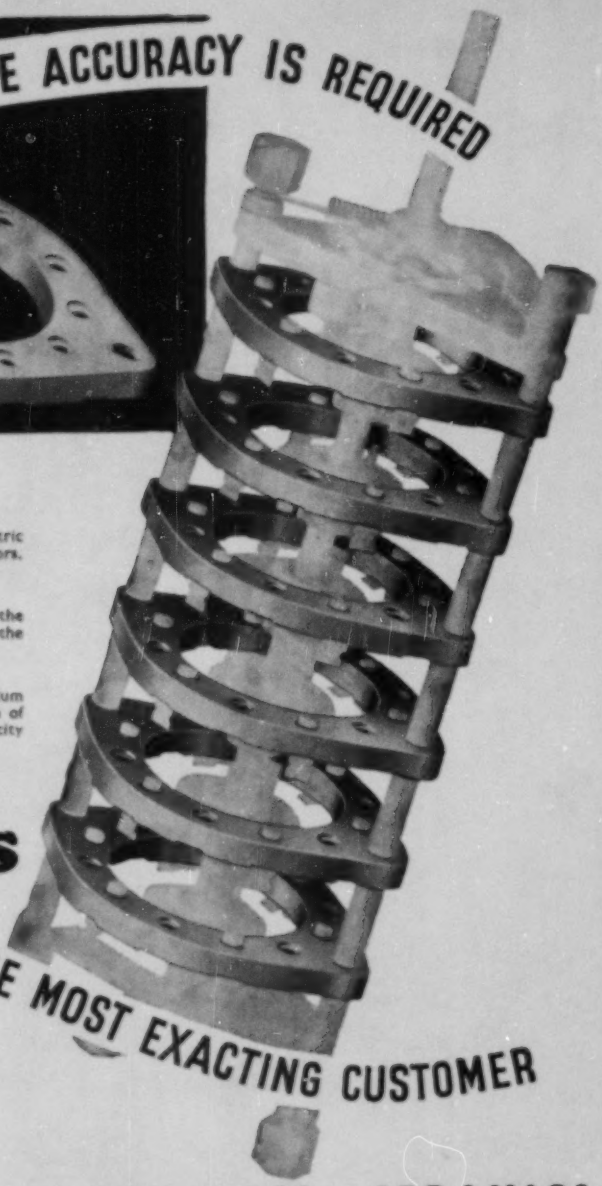
# WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

JULY 1950

VOL. XXVII TWO SHILLINGS AND SIXPENCE No. 322

# WHERE EXTREME ACCURACY IS REQUIRED



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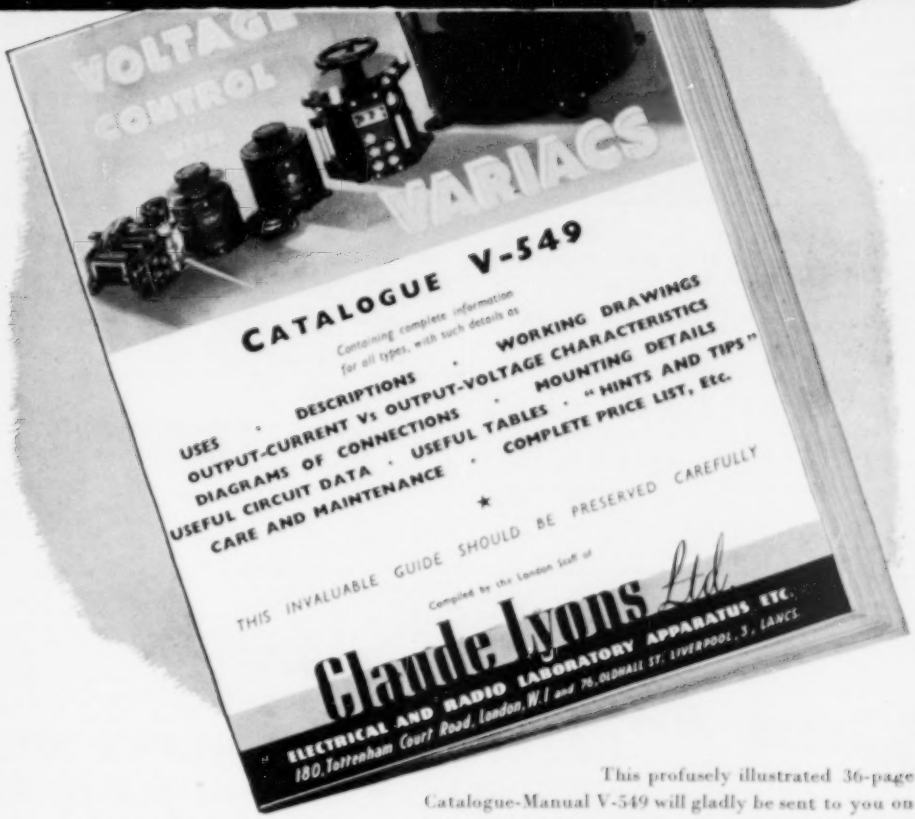
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CATALOGUE V-549
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for all types, with such details as
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D/22

dag DISPERSIONS

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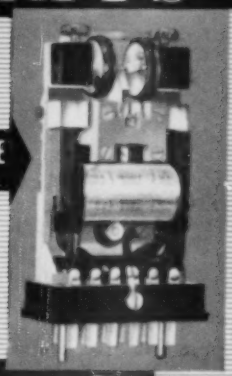
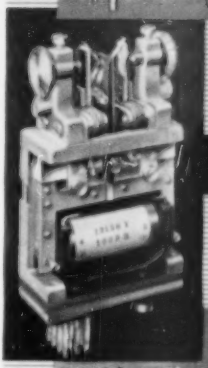
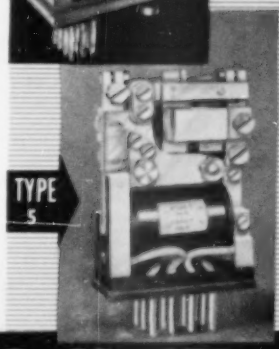
TMC

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**High operational speeds • Freedom from contact rebound and positional error • Good contact pressure • Accuracy of signal repetition • Sensitivity of the order of 0·12 Mw. D.C. • Ease of contact adjustment • Contact gap a function of input power.**

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**has longer contact travel than the Type 3, with medium speed and sensitivity. Also available in double-pole, double throw version with self-synchronising contacts.**
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*Note: All dimensions quoted exclude connection terminals.*

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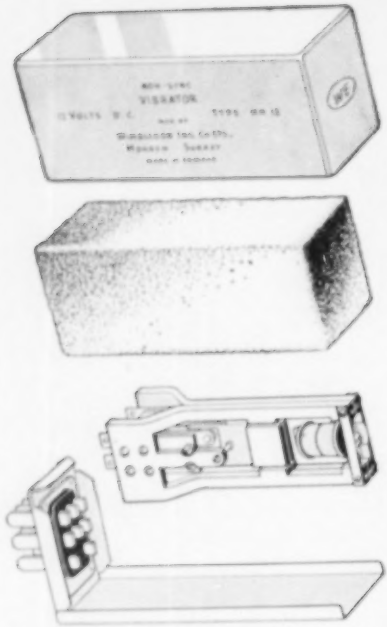

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CRC 4

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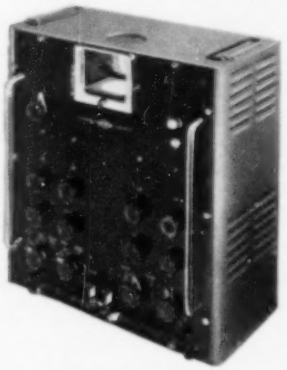

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RESISTANCE:
Model 1. 0.1 ohm to 5 megohms.
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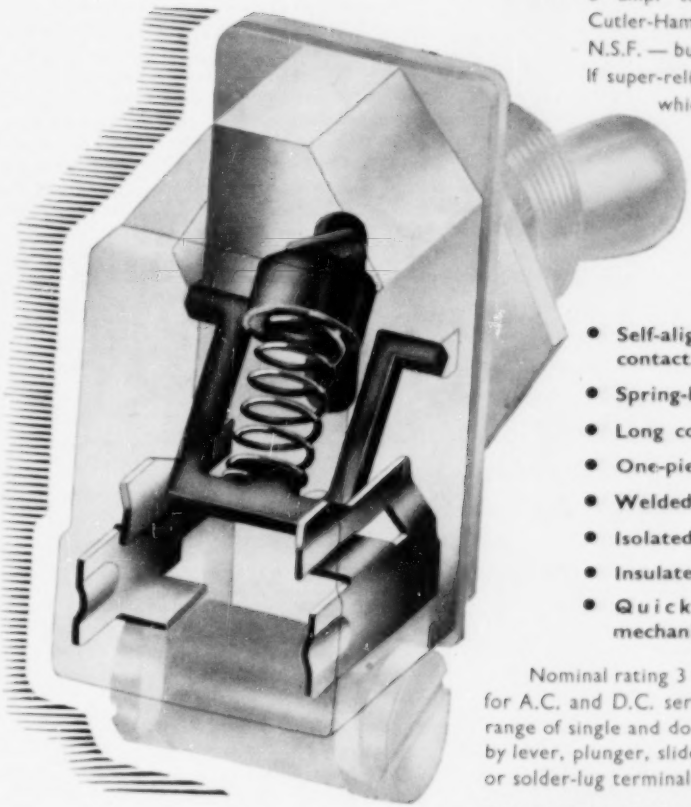
# INSIDE INFORMATION

## *about the switch that is superior*



We agree that there are cheaper 3 amp. toggle switches than the Cutler-Hammer types made by British N.S.F. — but there are none better. If super-reliability is the yardstick by which you choose a switch then this is the switch for you.

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**Other British N.S.F. Products include: 'OAK' Rotary, Push-button and Slider switches; Carbon and wire-wound potentiometers for Television and Radio applications.**

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R&A

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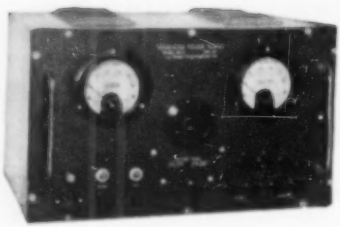

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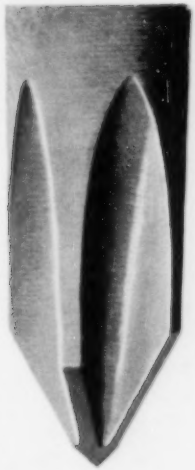
## WITH PHILLIPS RECESSED HEADS



*Available in a complete range of head styles*

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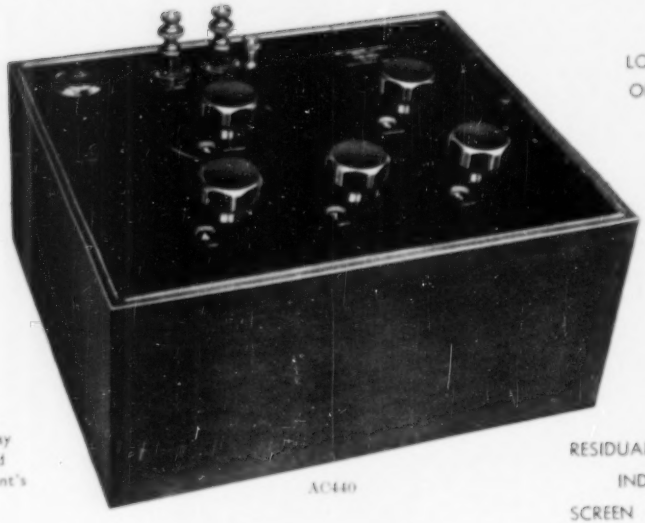

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# WIRELESS ENGINEER

**The Journal of Radio Research and Progress**



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## JULY 1950

Vol. XXVII No. 322


## CONTENTS


EDITORIAL. Surface-Wave Transmission Line .. .. .. 195

SIMPLE WAVE ANALYSER. By D. Martineau Tombs, M.Sc. 197

NEGATIVE-FEEDBACK AMPLIFIERS. By J. E. Flood, B.Sc. (Eng.) .. .. .. .. .. .. .. .. 201

INPUT IMPEDANCE OF TWO CROSSED DIPOLES. By L.L. G. Chambers, M.Sc. .. .. .. .. .. .. 209

DESIGN OF CATHODE-COUPLED AMPLIFIERS. By S. G. F. Ross, B.E. .. .. .. .. .. .. .. .. 212

CORRESPONDENCE .. .. .. .. .. .. .. .. 216

BOOK REVIEW .. .. .. .. .. .. .. .. 216

ABSTRACTS AND REFERENCES. Nos. 1560–1837 .. A.121–A.140



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EL91 Output Pentode (6AM5) (1.4W Output)
ECC91 V.H.F. Double Triode (6J6) (physical common cathode)
DK91 Vari-mu Heptode Frequency Changer (1R5)
DF91 Vari-mu R.F. or I.F. Pentode (1T4)
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DAF91 Single Diode Pentode (1S5)
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DL94 Output Pentode (3V4) (0.27W Output)
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DCC90 R.F. Double Triode (3A5)
2D21 Gas-filled Tetrode (Vh = 6.3V)

**A RANGE OF INDUSTRIAL PHOTOCELLS.** The Mullard range of all-glass photocells finds a particularly wide range of applications in industry and research. The types with caesium silver-oxide cathodes (type C), have a high sensitivity to incandescent light and near infra-red radiation and are particularly recommended for applications where a high output is required. For colour matching and photometric applications, photocells with caesium antimony cathodes (type A) are available, with a particularly high sensitivity to daylight and radiation throughout the visible colour spectrum. The end-view photocell 58CV, is intended for use in compact photometric units where conservation of space is of utmost importance.

| No. | Type | Cathode Surface | Base | Sensitivity (μA/l) |
|---|---|---|---|---|
| 20CV | Vacuum | C | B8G | 25 |
| 20CG | Gas-filled | C | B8G | 150 |
| 90CV | Vacuum | C | B7G | 20 |
| 90CG | Gas-filled | C | B7G | 125 |
| 20AV | Vacuum | A | B8G | 45 |
| 90AV | Vacuum | A | B7G | 45 |
| 58CV | Vacuum | C | Wire-in | 15 |
| 58CG | Gas-filled | C | Wire-in | 85 |

**MULLARD DISC-SEAL TRIODES FOR U.H.F.**

For applications such as U.H.F. links, local oscillators, and wide band applications such as multi-channel communication systems, radar altimeters, etc., the use of frequencies up to 1500Mc/s or even 3000Mc/s is often required. The Mullard Disc-Seal Triodes have been designed specifically for such purposes.

In the ME1001, adequate cooling and low inductance are combined in the design of the anode and grid discs to give an anode dissipation of 10 watts and an upper frequency limit (as an oscillator) of 3750-3500Mc/s (8-8.5cm).

The ME1003 is basically similar in design to the ME1001, but the larger electrodes permit an anode dissipation of 25 watts. However, the larger clearances necessary to obtain this value, limit the upper frequency to 2500Mc/s.

PRINCIPAL CHARACTERISTICS

| | ME1001 | ME1003 |
|---|---|---|
| Heater Voltage (V) | 6.3 | 6.3 |
| Heater Current (A) | 0.4 | 1.0 |
| Max. Anode Voltage (V) | 350 | 500 |
| Max. Anode Dissipation (W) | 10 | 25 |
| Max. Anode Current (mA) | 50 | 200 |
| Max. Peak Anode Current (mA) | 150 | 500 |
| Power Output (W) | 0.5 (at 3000 Mc/s) | 6 (at 1500 Mc/s) |
| | 3.5 (at 500 Mc/s) | 20 (at 430 Mc/s) |
| Mutual Conductance (mA/V) | 6 | 20 |
| Amplification Factor | 30 | 30 |

If you require full technical data on any of the valves described in this review, please check this form, pin to your letter heading and post to the address below.

Miniatures - - - - - ☐
Industrial Photocells - - ☐
Disc-Seal Triodes for U.H.F. ☐

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MVT 20

# WIRELESS ENGINEER

Vol. XXVII JULY 1950 No. 322


## Surface-Wave Transmission Line

OUR Editorial of June 1936 was entitled "New Type of Wave Transmission"; that was only fourteen years ago, and the new type was the tubular waveguide. In the opening paragraph we referred to the fact, which had been known for many years, that waves can be transmitted along a single wire. We also mentioned that the wire could be either a conductor or a dielectric. The single wire and the tubular waveguide are to some extent complementary, for, starting with a coaxial cable, if the inner conductor is removed, one has a tubular waveguide, whereas, if the outer conductor is removed, one has a single-wire transmission *line*. In both cases the electric field, instead of passing to the other conductor, curves round and returns to the same conductor. Although in recent years the tubular waveguide has made phenomenal progress and become a widely-used piece of apparatus, little use has been made of single-wire transmission. In the May number of the Radio-Electronic Engineering edition of *Radio & Television News* there is an article with the above title by Dr. Goubau of the American Signal Corps Engineering Laboratories, in which he describes some experiments made with such a transmission system.

Fig. 1 is a reproduction of Fig. 1 of our 1936 Editorial; it shows how the electric flux, which, in the presence of an outer conductor, would have been radial, curves round and returns to the conductor. An interesting question now arises as to the distribution of this field; how far do the electric lines of force extend before they curve round and return? As Dr. Goubau points out, it is important to keep the field as close as possible to the wire, both to avoid interference and as we shall see, to obtain a high efficiency of transmission. It is shown that the radial extent of the wave depends very largely on the phase velocity; a small decrease in the phase velocity, and therefore in the wavelength on the wire, causes an increased concentration of the field in the neighbourhood of the wire. The phase velocity will be reduced slightly by the resistivity of the conductor; at the frequencies employed skin effect will be very pronounced. The author also states that cutting a screw thread on the wire causes an appreciable reduction of the phase velocity; for the very thin effective conducting layer, the corrugation will lengthen the actual path of the surface currents. The method mainly employed, however, consists in coating the wire with enamel; even if the wire were a perfect conductor, a coating of enamel would cause a decrease in the phase velocity and a consequent spatial limitation of the field.

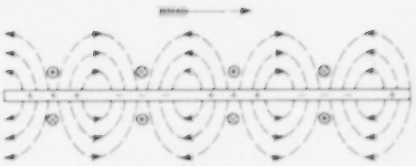

Fig. 1

The method employed by the author for setting up the wave consists in terminating the outer conductor of a coaxial cable in a funnel-shaped horn, while the inner conductor is continued as the transmission line. A similar horn at the receiving end collects the wave and carries it into a coaxial cable. This is illustrated in Fig. 2. The diameter of the mouth of the horn was 13 inches (not 13 feet as indicated in his diagram). Experiments were made with a 2-mm copper wire 120 ft long at frequencies of 1 600, 3 300, and 4 700 Mc/s. The author claims that the waves can be passed from the coaxial cable to the single line by means of the horn with a loss of only 10 per cent, but

the appropriate size of horn will depend on the wavelength. Unless the field is confined to the immediate neighbourhood of the wire, some of it will fail to pass into the collecting horn and thus reduce the efficiency. By the reciprocity

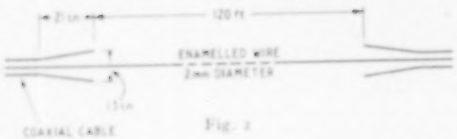


Fig. 2

theorem the same will be true at the transmitting horn. Fig. 3 shows the losses in decibels for the 120-ft line. In the calculations the enamel was assumed to have a dielectric constant of 3, a power-factor of 0.008, and a thickness of 0.05 mm. $L_c$, $L_h$ and $L_i$ are the losses in the conductor, horns, and insulation respectively; $L_{tot}$ is the total loss. It is stated that the losses in the same length of a very good cable at the same frequencies are from 6 to 10 times as great. It will be seen that the losses in the horns increase rapidly with the wavelength. Fig. 4 shows how the losses vary with the thickness of the dielectric coating. These curves are, we understand, based on calculations, but have been verified experimentally.

Particulars are given of another very striking experiment. Tests were made at a frequency of 2 600 Mc/s on an outdoor copper wire 2.6 mm diameter 120 ft long with the same horns as before. Having been outside for some months the wire was corroded, but on cleaning it with sandpaper *the loss went up from* 3 *to* 3.6 *db*; perhaps this is not surprising if the efficiency

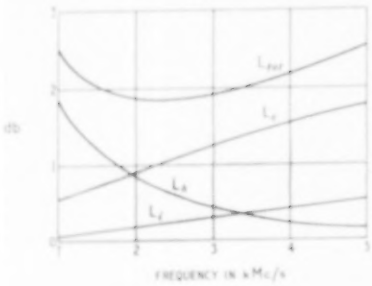


Fig. 3

can be increased by cutting a screw-thread on the wire. The wire was then given several coats of polystyrene solution, and the loss went down more and more until it reached a minimum of 1.7 db, but on adding further coats it went up to 1.9 db. The thickness was too variable for any value of the average thickness to be given. As we mentioned at the beginning the line can be either a conductor or a dielectric, and the results of these experiments seem to indicate that the layer of dielectric plays an important part in the transmission. It is true that it is very thin, but due to skin effect, the effective thickness of the conducting layer of the copper wire, at the frequencies employed is only about a thousandth of a millimetre, so that the transmission line may be regarded as a polystyrene cylinder with a metal lining.

Tests were also made on a 600 ft line of 3.2 mm copper wire with a layer of enamel 0.25 mm thick. The distance from the ground varied between 4 and 8 feet, and the wire was supported every 80 feet by waxed strings. The measured loss at 1 600 Mc/s was 5 db whereas the same length of a very good cable has an attenuation of 70 db. The effect of rain was investigated; a uniform film of water did not increase the loss, but rows of big rain drops did, the maximum increase being 1.5 db.

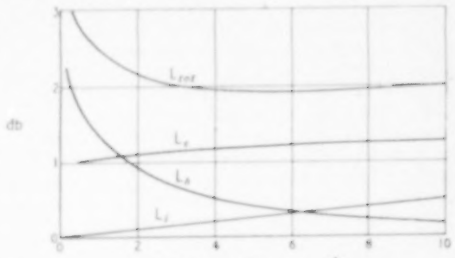


Fig. 4

The article does not show how the calculated results were arrived at, nor does it discuss in detail the phenomena involved. This will doubtless follow.

The discussion of this type of transmission in Stratton's "Electromagnetic Theory" shows that one cannot go far without becoming involved in very complex mathematics. The field distribution and the energy involved do not lend themselves to simple calculation, but the striking experimental results obtained by Dr. Goubau show that it is a subject that must be fully investigated.

G. W. O. H.

A slight reduction in the number of pages in *Wireless Engineer*, and some delay in publication, are still unavoidable on account of the continued withdrawal of overtime working by a section of the printing industry. All journals printed in London are similarly affected, to a greater or lesser extent, but journals printed in the provinces are unaffected.

# SIMPLE WAVE ANALYSER

By D. Martineau Tombs, M.Sc., A.C.G.I., D.I.C., A.M.I.E.E.

(Dept. of Electrical Engineering, Imperial College)

**SUMMARY.**—The device, in principle, uses the selective property of a simple tuned circuit to isolate a particular component of frequency existing in a complex wave.

The wave to be analysed is displayed on one trace of a double-beam oscilloscope and the selected component on the other. The amplitude of the harmonic being selected is governed by the selectivity (or $Q$), which can be adjusted over a wide range by the use of a negative-resistor. Under conditions of adequate selectivity, the $Q$ may then be determined by detuning. If, further, the $Q$s are adjusted successively to make the magnified voltage the same for two harmonics, the ratio of harmonic amplitudes is given by $e_m/e_n = (\omega_m/\omega_n)^2 (\Delta c_m/\Delta c_n)$, where $e_m$ = e.m.f. of the $m^{th}$ harmonic; $e_n$ = e.m.f. of the $n^{th}$ harmonic. The ratio $\omega_m/\omega_n$ comes from the display, the ratio $\Delta c_m/\Delta c_n$ from the half-power points. The use of a mutual-inductive coupling and a cathode follower extend the usefulness of the analyser.

The apparatus required for such a device is normally available in most laboratories, or can readily be made. The cost is a fraction of that of a commercial wave analyser.

## Introduction

THE circuit of Fig. 3 shows the essential parts of an early form of analyser. Its performance may best be described by reducing the network to an equivalent circuit. Thus, referring to Fig. 1(a), it will be seen that the analyser is basically a simple tuned circuit consisting of a fixed linear inductor $L$, the inductance of which is supposed constant in value (or known) over the frequency range covered by the analyser, a variable capacitor $C$, having a known incremental calibration, and which, with the inductor, allows tuning over the frequency range. $e$ represents the e.m.f. of the unbalanced source to be analysed, and $E_n$ the r.m.s. value of the magnified (or selected) voltage.

The process of analysis consists in deducing in turn the r.m.s. value of each component, $e_n$, present in the complex wave, measurement being made in the presence of all other components. This selection is done by reducing the effective series resistance $R$ until $E_n$ is purely sinusoidal, then measuring $E_n$ and the magnification, from which $e_n$ may be deduced.

The theory will be briefly stated on the assumption, which will be justified later, that adequate selectivity can be produced.

## Theory

First consider the circuit of Fig. 1(a) tuned to a particular component $e_n$, all other components being of negligible amplitude. Let $E_n$ be the maximum r.m.s. value of the voltage appearing across the capacitor, then

$$E_n = Qe_n \qquad \ldots\ldots \qquad (1)$$

where $Q$ is the magnification (or selectivity) of the circuit.

It is well known that $Q$, for reasonably high values, may be determined by detuning. If the capacitor is detuned by $\pm \Delta C/2$ from its tuned value $C_r$ until $E_n$ drops to $E_n/\sqrt{2}$

MS accepted by the Editor, December 1949

then $Q = 2C_r/\Delta C \qquad \ldots\ldots \qquad (2)$

From (1) and (2)

$$e_n = \frac{\Delta C}{2C_r} E_n \qquad \ldots\ldots \qquad (3)$$

And since at resonance

$$C_r = 1/\omega^2 L \qquad \ldots\ldots \qquad (4)$$

then $$e_n = \frac{\Delta C}{2} \omega^2 L E_n \qquad \ldots\ldots \qquad (5)$$

This is the basic equation for the equivalent circuit. It may be extended to cover the circuit of Fig. 1(c), Let $Y$ denote the unilateral transfer admittance of the cathode follower, and $M$ the mutual inductance between the small coil (in practice a few turns) in the cathode of the follower and the coil of the resonant circuit. If $Y$ be regarded as constant over the range of harmonic frequency involved, it is given by $i_n/V_n$ for each harmonic; or in general

$$Y = I/V \qquad \ldots\ldots \qquad (6)$$

thus $$V_n = \frac{\Delta C}{2} \omega \frac{1}{Y} \frac{L}{M} E_n \qquad \ldots\ldots \qquad (7)$$

The advantage of the magnetic coupling is that the induced e.m.f. increases with frequency, and this tends to offset the reduced amplitude of the higher-order harmonics that are usually encountered in practice.

The introduction of the cathode follower makes the input impedance of the device large, and its unilateral property isolates the source from the highly-selective tuned circuit. Harmonic distortion due to non-linearilty appears to be negligible in practice, as may be verified by a current analysis using the circuit of Fig. 1(b) with and without the cathode follower in circuit.

## Selectivity

In deriving the above equations it has been assumed that all harmonics except the one being studied have been sufficiently attenuated. This has implied an adequate selectivity in the resonant circuit.

Such a performance can be realized by the application to the circuit of one of a variety of known active networks. In the first models of the analyser a tetrode valve adjusted to the region of its negative slope was employed (Fig. 3), control being obtained by a variable resistor $R_a$ in the cathode.

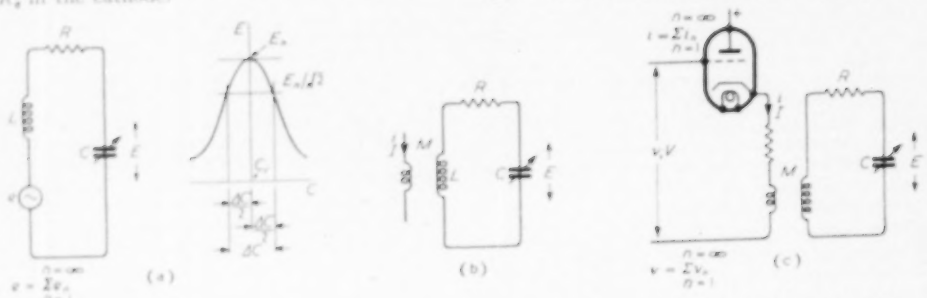


Fig. 1. *The equivalent circuits of simple types of analyser: (a) source e being analysed; (b) source I being analysed; (c) source V being analysed.*

Provided the amplitude of swing along the anode characteristic of the tetrode is maintained small (one volt r.m.s. in this case) the behaviour of the valve may be considered linear, and represented by a single symbol $r$* (Fig. 2.)

The effective selectivity $Q$ of the circuit shown in this figure may be calculated by considering the $LC$ circuit (the dynamic impedance $Z$ of which at its anti-resonant frequency is $L/CR$,) is in parallel with $r$. Thus for the circuit of Fig. 2

$$Q = \frac{1}{\omega L}\left[\frac{Z}{1 + Z/r}\right] \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

As $r$ is reduced from a high initial negative value, $Q$ increases continuously. Stable values of several hundred may be realized at frequencies as low as a few hundred cycles per second, and it is not difficult in practice to satisfy the requirement of adequate reduction of all other harmonic components relative to the one under consideration. Instability occurs when $r$ becomes less in magnitude than $Z$.

## Tests for Adequate Selectivity

If the selectivity is adequate, the selected voltage $E_n$ will be purely sinusoidal. The display on an oscilloscope will not, however, give more than a rough indication of the purity of the wave, and a more rigid test is necessary. It is evident from the foregoing discussion, that tests leading to the same results may be made at any suitable selectivity provided it is sufficiently high. Hence by successively increasing the selectivity an asymptotic value, which will be the true one, will be reached. Thus $e_n$ of Fig. 1(a), $i_n$ of Fig. 1(b) or $v_n$ of Fig. 1(c) are accurately determined.

Nevertheless, by examining the waveforms to be expected from an adjustment with insufficient selectivity, it may be shown that if $n$ is the order of harmonic tuned to, and $m$ the order interfering (or breaking through), then the peaks of the displayed wave will be non-uniform in height if $m < n$ and also for all values of $m > n$ except $m = (2, 3, 4 \ldots) n$. Thus, if the third harmonic were tuned to ($n = 3$), a strong sixth harmonic ($m = 6$) would not alter the relative height of successive peaks, and thus would constitute a failure of a test for adequate selectivity based on the uniformity of peaks. In practice, however, the test of uniformity of peak height enables the circuit to be rapidly adjusted to the desired selectivity, the final necessary step being a check at a slightly higher degree of selectivity.

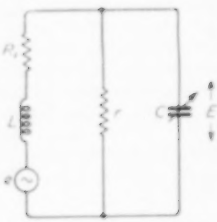


Fig. 2. *The addition of a voltage - controlled negative resistance r to reduce the effective losses.*

It should be noted that the worst failure occurs with the fundamental, since the presence of harmonics in this case does not alter the magnitude of successive peaks. The difficulty, however, is not as serious as might be supposed, since the fundamental is more widely separated from the second harmonic than any other harmonic from its adjacent one.

Thus, in practice, the first test, and often a sufficient one, is to increase the selectivity until the peaks are of equal height. The second test is to make a further measurement with a greater selectivity. If the results are sufficiently close a third test at still higher selectivity is unnecessary.

* To avoid confusion about signs, the symbol $r$ is taken to mean a resistance of positive or negative value. In this application the dynatron had a minimum value when negative of $r = -23{,}000\Omega$.

Fig. 4 shows a series of oscillograms taken using a double-beam oscillograph. The wave being analysed was kept on the upper trace as a reference system, while the tips of the selected harmonic were displayed on the lower trace. The separate oscillograms were taken when adequate selectivity had been reached at each hormonic in turn. It will be noted that not only the amplitudes of different selected harmonics (on different pictures) are adjusted to the same value, but that the individual peak heights of harmonics within the period of a cycle of the fundamental, are also made the same. The technique of doing this is described later.

It may be objected that on close examination of the harmonic trace, particularly of the right-hand group, the individual amplitude of peaks is not constant, indicating insufficient selectivity. This, in fact, is not the case, since the inequality is due to interbeam coupling within the cathode-ray tube itself, a source of error quickly recognized by a reduction of the reference-trace amplitude.

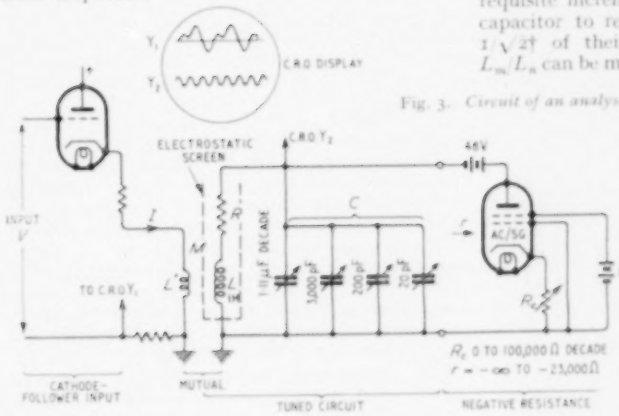


Fig. 3. *Circuit of an analyser.*

## Stability Criteria

Although, in equation (8) it has been considered adequate to take the negative-resistance as linear, at the threshold of oscillation the second-order curvature is of importance if the device is to move smoothly and not discontinuously into and out of oscillation under the control of $R_c$ (Fig. 3). If, for an increasing amplitude of signal about the operating point, " the effective negative-resistance " becomes greater, the circuit is stable, but if the reverse is the case, discontinuous operation occurs, and the amplitude of the display increases indefinitely, no longer under the incremental control of $R_c$. $r$ must be very greatly increased by increasing $R_c$ and the condition of high selectivity re-approached slowly by reducing $r$. This may be given mathematical formulation by stating that if the dynatron characteristic is plotted as $I = f(V)$ the circuit in the region of negative resistance is stable if $d^2I/dV^2$ is positive and unstable if negative.

If the wave to be analysed is displayed on one trace of a double-beam oscilloscope, and the analysed component on the other, the condition of oscillation is immediately evident from relative movement of the two patterns.

## Comparing Harmonics

If two frequency components, $m$ and $n$, are to be compared in amplitude, then, using equation (7).

$$\frac{V_m}{V_n} = \frac{Y_n}{Y_m} \frac{E_m}{E_n} \frac{\omega_m}{\omega_n} \frac{\Delta C_m}{\Delta C_n} \frac{L_m}{L_n} \frac{M_n}{M_m} \quad \ldots (9)$$

and the analysis is performed when the above series of ratios is determined.

Of these ratios, $\omega_m/\omega_n$, comes directly from the display by counting, and $\Delta C_m/\Delta C_n$ from the requisite incremental adjustments of the tuning capacitor to reduce $E_m$ and $E_n$ respectively to $1/\sqrt{2}$† of their resonant values. $Y_m/Y_n$ and $L_m/L_n$ can be made very close to unity. Likewise $E_m/E_n$ may be made unity in two ways, as follows: if a standard mutual inductance is available, the calibration of which is valid over the frequency range, convenient known values of $M$ may be used for the different harmonics. By such means $E_m$ may be made equal to $E_n$, leaving the ratio $M_n/M_m$ in the equation. Failing such a mutual inductance, $M$ is kept unchanged ($M_n/M_m = 1$), and $E_m$ and $E_n$ made equal by suitable use of the variable negative resistance $r$.

As discussed above $r$ must be adjusted to give at least sufficient selectivity, but there is no objection to increasing the selectivity further, where necessary. This freedom to increase the selectivity allows the observed value of $E$ on the oscilloscope to be varied arbitrarily. A convenient technique is to choose a suitable amplitude on the oscilloscope and adjust the selected harmonics in turn to this value, if necessary returning over earlier measurements which may have to be repeated with a higher degree of selectivity to meet the dual requirements of adequate selectivity and equality of $E_m$ and $E_n$. It is obviously undesirable to use a greater degree of selectivity

† Or to any other convenient ratio.

than is necessary since the circuit is proceeding towards instability.

A direct, though approximate, measure of the actual value of $E$ is given from the oscilloscope trace, but the measure may be obtained alternatively by use of a valve voltmeter, provided care is taken to ensure that the input impedance of the voltmeter does not introduce errors at high $Q$ values.‡ The advantage of adjusting the deflection to a constant value avoids the need of a calibrated instrument.

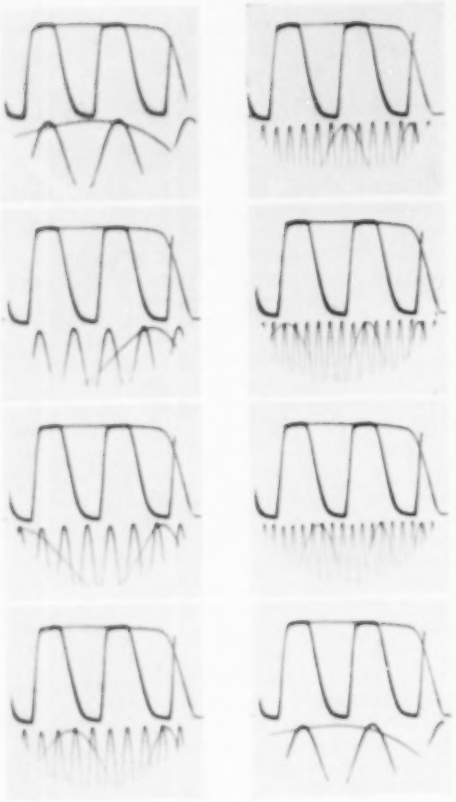

Fig. 4. *A series of harmonics recorded during a test. The upper trace is of the wave being analysed, while the lower trace shows the peaks of selected harmonics.*

‡ The dynamic impedance of the circuit at such values may be several megohms, and is directly given by $E = I/(\omega dC)$. This provides a convenient method of measuring the input impedance of such devices.

## Experimental Results

In Table 1 the results of comparative analysis as carried out by the method described, and by the use of a good commercial analyser, are shown. In this case the source being examined was a fork-controlled oscillator. A high degree of frequency stability is essential if very selective circuits are to be employed. Fig. 4 shows a series of oscillograms taken during a test, and employing the technique of the previous section.

TABLE 1

| Harmonic Number $n$ | Author's Analyser | | Commercial Analyser |
|---|---|---|---|
| | $\omega C_n$ ($\mu$F) between the two $\frac{1}{2}$-power points | Harmonic Ratio $n\frac{\Delta C_n}{\Delta C_1} \times 100\%$ | Harmonic Ratio % |
| 1 | 14,000 | 100 | 100 |
| 2 | 910 | 13.0 | 13.7 |
| 3 | 721 | 15.4 | 16.3 |
| 4 | 210 | 6.0 | 6.0 |
| 5 | 103 | 3.89 | 4.0 |
| 6 | 54.6 | 2.3 | 2.4 |
| 7 | not measured | | |
| 8 | | | |
| 9 | 6.56 | 0.42 | 0.4 |

## Stray Fields

It is surprising at first, in using these highly-selective circuits, to find how sensitive they are to stray fields in a laboratory. Tuning to the seventh harmonic of the mains and moving the coil $L$ into different positions, the field pattern both in magnitude and direction, due to a piece of mains-operated apparatus can readily be mapped. It is also helpful for demonstration purposes to couple the output of the selected harmonic on to a loud-speaker, thus facilitating the exploration of stray fields where the oscilloscope may not be visible.

The apparatus substantially as described herein, was exhibited at the Centenary Celebrations at Imperial College in November 1945, and subsequently at the Physical Society's Exhibition in January 1946. Since then, W. I. Heath and J. A. Fredericks, at the General Electric Research Laboratories, Wembley, England, have made a mains-operated analyser dependent on the above principle. The first G.E.C. model was shown at the Physical Society's Exhibition in April 1949, on the author's stand.


## Acknowledgment

The author wishes to thank Professor Willis Jackson and Mr. M. F. McKenna for helpful criticism in the preparation of the paper.

# NEGATIVE-FEEDBACK AMPLIFIERS

*Conditions for Critical Damping*

By J. E. Flood, B.Sc. (Eng.)

## 1. Introduction

WHEN negative feedback is applied to a resistance-coupled amplifier having two or more stages it is well-known that, once a certain amount of feedback is exceeded, the frequency characteristic develops peaks at the edges of the band and the transient response of the amplifier contains a damped oscillation. Brockelsby[1] and Mayr[2] have shown that a substantially flat response over the maximum possible frequency band can be obtained by arranging for a suitable relationship between the time constants of the stages of the amplifier. The condition of maximal flatness, however, still gives an oscillatory transient response: for instance, when the input voltage is a unit step (i.e., a voltage which is zero when $t<0$ and unity when $t>0$) the two-stage maximally-flat amplifier has a response which overshoots 4.3%, and the three-stage amplifier overshoots 8%.

The present analysis determines the relationship required between the time constants of a two-stage amplifier with a feedback path whose transfer coefficient $\beta$ is independent of frequency in order to obtain the most rapid transient response which is possible without overshoot; this corresponds to the condition of critical damping. It is shown that a modification of the feedback path enables a critically-damped response to be obtained when the ratio between the time constants of the stages is less than that required for critical damping when $\beta$ is independent of frequency. It is shown that a critically-damped transient response can also be obtained from a three-stage amplifier by modifying the feedback path.

The effective feedback of the amplifier decreases at high frequencies and at the high-frequency end of the band there is little or no feedback. The amplifier is therefore unsuitable for a multichannel carrier-system amplifier, but is suitable for video-frequency or pulse amplification. The steady-state gain of the amplifier is stabilized by the full amount of the feedback at low frequencies; the feedback at the higher frequencies only ensures that the bandwidth is adequate for a sufficiently rapid transient response and that the shape of the frequency-response curve is such that the transient response is free from ringing. The analysis assumes that the operation of the amplifier is linear. It is important that a signal shall not be applied to the amplifier which rises to sufficient magnitude to overload the amplifier before the voltage fed back has built up to its steady-state value.

MS accepted by the Editor, January 1950

## 2. Indicial Response of Feedback Amplifiers

If the Heaviside operational expression for the indicial response (i.e., the response to a unit step) of an amplifier is $\mu(p)\mathbf{1}$, and the input voltage is $e(t)\mathbf{1}$, the output voltage is

$$\mu(p).\ e(t)\mathbf{1}$$

Let the indicial response of the feedback path be $\beta(p)\mathbf{1}$, so that the voltage returned to the input circuit to produce feedback is

$$\mu(p).\ \beta(p).\ e(t)\mathbf{1}$$

If the input voltage to the feedback amplifier is a unit step, then

$$e(t)\mathbf{1} = \mathbf{1} - \mu(p).\ \beta(p).\ e(t)\mathbf{1}$$

$$\therefore\ e(t)\mathbf{1} = \frac{1}{1+\mu(p).\ \beta(p)}\mathbf{1}$$

$\therefore$ The output voltage is

$$h(p)\mathbf{1} = \frac{\mu(p)}{1+\mu(p).\ \beta(p)}\mathbf{1} \qquad \ldots\ \ldots\ (1)$$

which is the indicial response of the feedback amplifier. The sign convention is chosen so that $\mu$ and $\beta$ are both positive quantities when the feedback is negative.

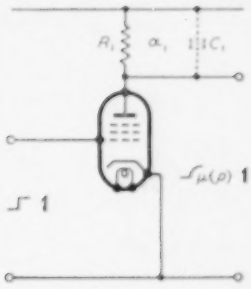


Fig. 1. *Basic RC amplifier circuit.*

## 3. Single-stage Amplifier

The type of amplifier considered is shown in Fig. 1. The anode load resistance is $R_1$ and $C_1$ is the total stray capacitance. If the steady-state low-frequency gain of the amplifier without feedback is $\mu_0$, its indicial response is

$$\mu(p)\mathbf{1} = \mu_0\frac{\alpha_1}{p+\alpha_1}\mathbf{1} = \mu_0(1-e^{-\alpha_1 t})\mathbf{1}$$

where $\alpha_1 = 1/R_1C_1$

If the transfer coefficient of the feedback path is $\beta_0$ and is independent of frequency, the indicial response of the amplifier with feedback is

$$h(p)\mathbf{1} = \frac{\mu_0 \frac{\alpha_1}{p+\alpha_1}}{1+\mu_0\beta_0\frac{\alpha_1}{p+\alpha_1}}\mathbf{1}$$

$$= \frac{\mu_0\alpha_1}{p+\alpha_1(1+\mu_0\beta_0)}\mathbf{1}$$

$$= \frac{\mu_0}{1+\mu_0\beta_0}\cdot\frac{\alpha}{p+\alpha}\mathbf{1} \quad .. \quad .. \quad (2)$$

where $\alpha = \alpha_1(1+\mu_0\beta_0)$

$$\therefore h(t) = \frac{\mu_0}{1+\mu_0\beta_0}(1-e^{-\alpha t}),\ t>0 \quad .. \quad (3)$$

The feedback reduces the time constant of the amplifier in the same ratio as it reduces the gain at low frequencies. The steady-state gain is stabilized by the feedback so that if $\mu_0\beta_0 \gg 1$, it will remain almost constant when the gain of the stage decreases (e.g., due to valve aging), but the time-constant of the circuit increases as the gain decreases. In order to obtain a stable gain under pulse conditions it is therefore necessary that the time-constant with feedback shall be sufficiently small for the pulse response to reach substantially steady-state when the gain is the smallest value likely to be encountered in practice.

## 4. Two-stage Amplifier

Fig. 2 shows a typical two-stage amplifier and the three-stage amplifier shown in Fig. 3 also contains two time-constants within the feedback loop.

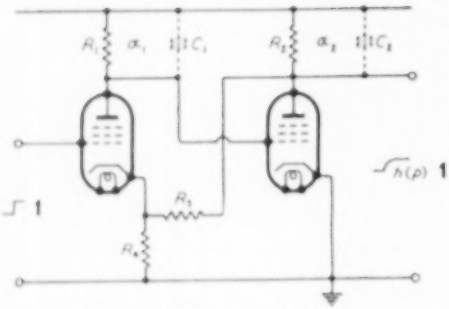


Fig. 2. *Two-stage feedback amplifier.*

The indicial response of the amplifying path is

$$\mu(p)\mathbf{1} = \mu_0\frac{\alpha_1}{p+\alpha_1}\cdot\frac{\alpha_2}{p+\alpha_2}\mathbf{1} \quad .. \quad .. \quad (4)$$

where $\alpha_1 = 1/R_1C_1$ and $\alpha_2 = 1/R_2C_2$.

The indicial response with feedback (the transfer coefficient being $\beta_0$ and independent of frequency) becomes

$$h(p)\mathbf{1} = \frac{\mu_0\frac{\alpha_1}{p+\alpha_1}\cdot\frac{\alpha_2}{p+\alpha_2}}{1+\mu_0\beta_0\frac{\alpha_1}{p+\alpha_1}\cdot\frac{\alpha_2}{p+\alpha_2}}\mathbf{1}$$

$$= \frac{\mu_0\alpha_1\alpha_2}{p^2+p(\alpha_1+\alpha_2)+\alpha_1\alpha_2(1+\mu_0\beta_0)}\mathbf{1} \quad .. \quad .. \quad (5)$$

$$= \frac{\mu_0\alpha_1\alpha_2}{(p+\lambda_1)(p+\lambda_2)}\mathbf{1}$$

where $(p+\lambda_1)$, $(p+\lambda_2)$ are the factors of the denominator of equation (5). If $\lambda_1$, $\lambda_2$ are complex, the response is of the form

$$h(t) = 1 - Ae^{-\alpha t}\sin(\omega t+\phi) \quad .. \quad .. \quad (6)$$

which represents a damped oscillation.

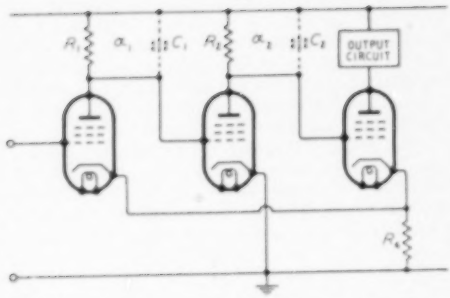


Fig. 3. *Three-stage feedback amplifier with only two time constants within the feedback loop.*

If $\lambda_1$, $\lambda_2$ are real, but unequal, the response is of the form

$$h(t) = 1 - Be^{-\alpha t}\sinh(\beta t+\gamma) \quad .. \quad .. \quad (7)$$

which is aperiodic.

If $\lambda_1 = \lambda_2 = \alpha$, the response is of the form

$$h(t) = 1 - (1+\alpha t)e^{-\alpha t} \quad .. \quad .. \quad (8)$$

which corresponds to critical damping and gives the quickest build-up which is possible without overshoot. Now

$$\lambda_1.\lambda_2 = \tfrac{1}{2}\{(\alpha_1+\alpha_2) \mp \sqrt{(\alpha_1+\alpha_2)^2 - 4\alpha_1\alpha_2(1+\mu_0\beta_0)}\}$$

$\therefore$ In order to ensure real factors

$$1+\mu_0\beta_0 \leqslant \frac{(\alpha_1+\alpha_2)^2}{4\alpha_1\alpha_2}$$

and when $1+\mu_0\beta_0 = \dfrac{(\alpha_1+\alpha_2)^2}{4\alpha_1\alpha_2}$,

the factors are equal and the overall indicial response is

$$h(p)\mathbf{1} = \frac{\mu_0 \alpha_1 \alpha_2}{(p+\alpha)^2}\mathbf{1}$$

$$= \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{\alpha^2}{(p+\alpha)^2}\mathbf{1} \quad .. \quad .. \quad (9)$$

where

$$\alpha = \tfrac{1}{2}(\alpha_1 + \alpha_2) = \sqrt{\alpha_1\alpha_2(1+\mu_0\beta_0)} \quad .. \quad (10)$$

$$\therefore h(t) = \frac{\mu_0}{1+\mu_0\beta_0}\,[1 - e^{-\alpha t}(1+\alpha t)],\ t > 0$$
$$.. \quad .. \quad .. \quad (11)$$

Fig. 4 shows the feedback factor $(1+\mu_0\beta_0)$ and the resulting time constant plotted against the ratio of the time constants of the two stages. If a large amount of feedback is to be used, the ratio of the time constants must be large.

The condition for maximal flatness (a term introduced by Landon[3]) is given by Brockelsby[4] as

$$F = 2S^2$$

where

$$F = (1+\mu_0\beta_0) \quad \text{and} \quad S = \frac{\alpha_1 + \alpha_2}{2\sqrt{\alpha_1\alpha_2}}$$

$$\therefore 1 + \mu_0\beta_0 = \frac{(\alpha_1+\alpha_2)^2}{2\alpha_1\alpha_2}$$

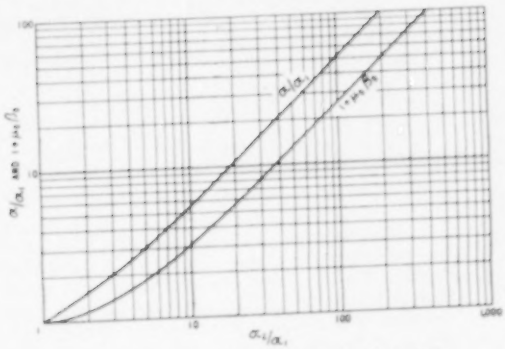


Fig. 4. *Conditions for critical damping of two-stage uncompensated amplifier.*

It is seen that for any particular ratio of $\alpha_2/\alpha_1$, the condition of critical damping permits a feedback factor of only one-half that required for maximal flatness of the frequency characteristic. The transient response of the maximally-flat amplifier is therefore oscillatory.

More feedback can be obtained by modifying the feedback path so that its indicial response is

$$\beta_0 \frac{p+\alpha_3}{\alpha_3}\mathbf{1}.$$

If a capacitance $C_3$ is connected in parallel with $R_3$, as shown in Fig. 5, and $R_3 \gg R_4$, then

$$\beta(p)\mathbf{1} \approx \beta_0(1 + pC_3R_3)\mathbf{1}$$

$$= \beta_0 \frac{p+\alpha_3}{\alpha_3}\mathbf{1}$$

where $\alpha_3 = 1/R_3C_3$

If an inductance $L$ is connected in series with $R_4$, as shown in Fig. 6, and $g_mR_4 \ll 1$ then

$$\beta(p)\mathbf{1} \approx \beta_0\left(1 + p\frac{L}{R_4}\right)\mathbf{1}$$

$$= \beta_0 \frac{p+\alpha_3}{\alpha_3}\mathbf{1}$$

where $\alpha_3 = R_4/L$.

The indicial response $\frac{p+\alpha_3}{\alpha_3}\mathbf{1}$ corresponds to an amplitude characteristic which rises with frequency and becomes infinite at infinite frequency. The amplitude characteristics of the $\beta$ paths of Figs. 5 and 6 cannot rise above unity, but the circuits approximate sufficiently well to the required response provided that $R_3 \gg R_4$ and $g_mR_4 \ll 1$.

The overall indicial response is

$$h(p)\mathbf{1} = \frac{\mu_0 \dfrac{\alpha_1}{p+\alpha_1} \cdot \dfrac{\alpha_2}{p+\alpha_2}}{1 + \mu_0\beta_0 \dfrac{\alpha_1}{p+\alpha_1} \cdot \dfrac{\alpha_2}{p+\alpha_2} \cdot \dfrac{p+\alpha_3}{\alpha_3}}\mathbf{1}$$

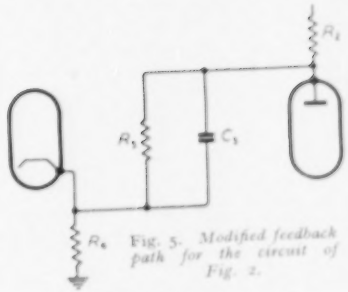


Fig. 5. *Modified feedback path for the circuit of Fig. 2.*

$$= \frac{\mu_0\alpha_1\alpha_2}{p^2 + p\left(\alpha_1 + \alpha_2 + \dfrac{\alpha_1\alpha_2}{\alpha_3}\mu_0\beta_0\right) + \alpha_1\alpha_2(1+\mu_0\beta_0)}\mathbf{1}$$
$$.. \quad .. \quad .. \quad (12)$$

The factors of the denominator are

$$(p+\lambda_1),\ (p+\lambda_2)$$

where

$$\lambda_1, \lambda_2 = \tfrac{1}{2}\left\{\left(\alpha_1 + \alpha_2 + \frac{\alpha_1\alpha_2}{\alpha_3}\mu_0\beta_0\right) \mp \sqrt{\left(\alpha_1 + \alpha_2 + \frac{\alpha_1\alpha_2}{\alpha_3}\mu_0\beta_0\right)^2 - 4\alpha_1\alpha_2(1+\mu_0\beta_0)}\right\}$$

If the factors are equal,

$$\alpha_2 = \frac{\alpha_1\alpha_2\mu_0\beta_0}{2\sqrt{\alpha_1\alpha_2(1+\mu_0\beta_0)} - (\alpha_1+\alpha_2)} \quad .. \quad (13)$$

and

$$h(p) \rfloor = \frac{\mu_0\alpha_1\alpha_2}{(p+\alpha)^2} \rfloor = \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{\alpha^2}{(p+\alpha)^2} \rfloor \quad (9)$$

where

$$\alpha = \sqrt{\alpha_1\alpha_2(1+\mu_0\beta_0)} \quad .. \quad .. \quad .. \quad (14)$$

$$\therefore h(t) = \frac{\mu_0}{1+\mu_0\beta_0}\left[1 - e^{-\alpha t}(1+\alpha t)\right], \; t>0 \; .. \quad (11)$$

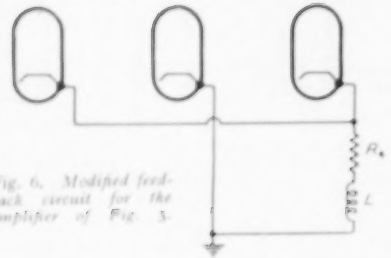


*Fig. 6. Modified feedback circuit for the amplifier of Fig. 3.*

By the choice of the value given by equation (13) for the time constant for the feedback circuit, critical damping is obtained and the indicial response of the feedback amplifier is made identical with that of an amplifier without feedback having two stages, each having a time constant equal to the geometric mean of the time constants of the two stages multiplied by the square root of the feedback factor. A particularly simple special case is obtained when $\alpha_2 = \alpha_1(1+\mu_0\beta_0)$, then $\alpha = \alpha_3 = \alpha_2 = \alpha_1(1+\mu_0\beta_0)$.

If the time constant of the anode load of the third stage of the amplifier of Fig. 3 is made $1/\alpha$, the indicial response of the three-stage amplifier is

$$h(p) \rfloor = \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{\alpha^3}{(p+\alpha)^3} \rfloor \quad .. \quad .. \quad (15)$$

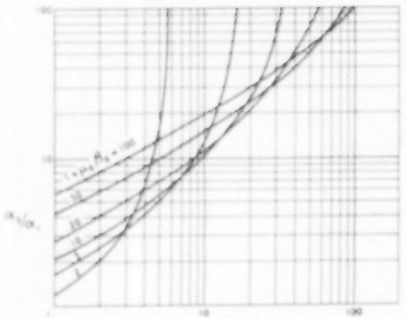


*Fig. 7. Conditions for critical damping of two-stage compensated amplifier.*

$$\therefore h(t) = \frac{\mu_0}{1+\mu_0\beta_0}\left[1 - e^{-\alpha t}\left(1+\alpha t + \frac{\alpha^2 t^2}{2}\right)\right], \quad t>0 \quad .. \quad (16)$$

Figs. 7 and 8 show the variation of the time constant of the amplifier and of the feedback path plotted against the ratio between the time constants of the two stages for different values of $1+\mu_0\beta_0$. Both $\alpha$ and $\alpha_3$ increase with an increase in this ratio until

$$1+\mu_0\beta_0 = \frac{(\alpha_1+\alpha_2)^2}{4\alpha_1\alpha_2},$$

when the time constant of the feedback path is zero. Any further increase in the ratio $\alpha_2/\alpha_1$ results in the amplifier having an overdamped response of the form of equation (7). If an increase is made in the gain $\mu_0$ of an amplifier which is critically-damped, the transient response will become oscillatory, and if the gain is reduced the response will become overdamped and less rapid.

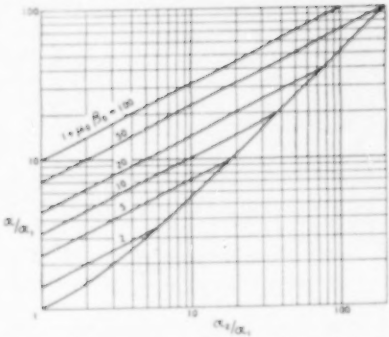


*Fig. 8. Time-constant of critically-damped two-stage compensated amplifier.*

It is interesting to examine the frequency characteristics of the two-stage amplifier circuit. Fig. 9 shows the amplitude and phase characteristics of an amplifier having the time constant of one stage ten times that of the other. Fig. 10 shows the corresponding Nyquist diagram when there is 26-db of feedback at low frequencies. When the locus of $\mu\beta$ is within the unit circle centre $(-1, 0)$ the feedback has become positive[4]. Although at frequencies at which the feedback is positive, the gain of the $\mu$ path has fallen, the net result is a peak in the overall amplitude characteristic which is shown in Fig. 11.

In order to avoid this undesirable rise in the amplitude characteristic, it is necessary to reduce the phase angle of $\mu\beta$ at high frequencies. The vertical line AA' on the Nyquist diagram is the locus for constant gain[4] and it is undesirable that the locus of $\mu\beta$ could cross this line. A

simple method of reducing the phase shift at high frequencies is to modify the feedback path of the amplifier as shown in Figs. 5 and 6.

Fig. 11 also shows as a broken line the amplitude characteristic of the amplifier when the feedback path is modified to give critical damping ($\alpha_3 = 1.16\alpha_2$). Fig. 10 shows as a broken line the Nyquist diagram corresponding to critical

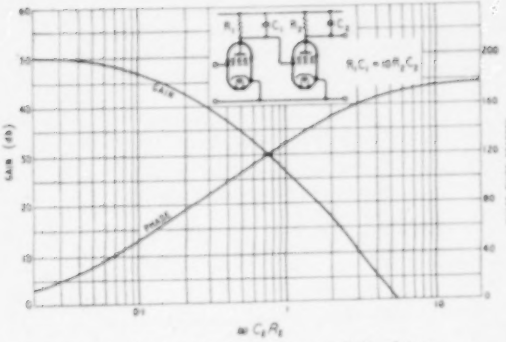


Fig. 9. *Amplitude and phase characteristic of two-stage amplifier with one time constant 10 times the other.*

damping and shows the large increase in the phase margin[5] which has resulted.

The gain of the amplifier without feedback is, from equation (4)

$$\mu(\omega) = \mu_0 \frac{\alpha_1 \alpha_2}{(\alpha_1 + j\omega)(\alpha_2 + j\omega)}$$

$$\therefore |\mu(\omega)| = \mu_0 \frac{\alpha_1 \alpha_2}{\sqrt{\alpha_1^2\alpha_2^2 + \omega^2(\alpha_1^2 + \alpha_2^2) + \omega^4}}$$

The overall gain of the critically-damped feedback amplifier is, from equation (9)

$$A(\omega) = \frac{\mu_0}{1 + \mu_0\beta_0} \cdot \frac{\alpha^2}{(\alpha + j\omega)^2} = \frac{\mu_0\alpha_1\alpha_2}{(\alpha + j\omega)^2}$$

where $\alpha = \sqrt{\alpha_1\alpha_2(1 + \mu_0\beta_0)}$

$$\therefore |A(\omega)| = \frac{\mu_0\alpha_1\alpha_2}{\alpha_2 + \omega_2}$$

At low frequencies $|A(\omega)| = \frac{1}{1 + \mu_0\beta_0} |\mu(\omega)|$

and at high frequencies $|A(\omega)| \approx \frac{\mu_0\alpha_1\alpha_2}{\omega^2} \approx |\mu(\omega)|$,

therefore the overall gain-frequency characteristic of the critically-damped amplifier is below that of its $\mu$ path at low frequencies and asymptotic to it at high frequencies. It can be simply shown that if

$$1 + \mu_0\beta_0 \geqslant \frac{\alpha_1^2 + \alpha_2^2}{2\alpha_1\alpha_2}$$

the overall gain-frequency characteristic is below that of the $\mu$ path at all frequencies and the locus of $\mu\beta$ is entirely outside the unit circle centre $(-1, 0)$, touching it at infinite frequency; i.e., the feedback is negative at all frequencies.

If $1 + \mu_0\beta_0 < \frac{\alpha_1^2 + \alpha_2^2}{2\alpha_1\alpha_2}$ the overall gain-frequency characteristic cuts that of the $\mu$ path and is above it at high frequencies, and the locus of $\mu\beta$ cuts the unit circle centre $(-1, 0)$ and is within it at high frequencies; i.e., the feedback is positive at high frequencies. In particular, if $\beta$ is independent of frequency the critically-damped amplifier has positive feedback at high frequencies and if $\alpha_1 = \alpha_2$ the feedback is always negative. For the example discussed above ($\alpha_2 = 10\alpha_1$, $\mu_0\beta_0 = 20$), when the amplifier is critically damped the feedback is negative at all frequencies, as shown by Figs. 10 and 11.

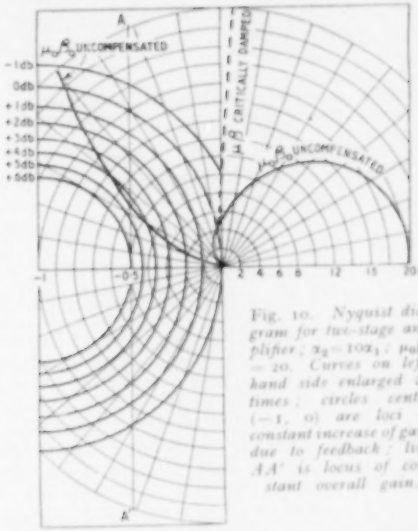


Fig. 10. *Nyquist diagram for two-stage amplifier; $\alpha_2 = 10\alpha_1$; $\mu_0\beta_0 = 20$. Curves on left-hand side enlarged 20 times; circles centre $(-1, 0)$ are loci of constant increase of gain due to feedback; line $AA'$ is locus of constant overall gain.*

## 5. Three-stage Amplifier

If an amplifier has three stages with time constants $1/\alpha_1$, $1/\alpha_2$, $1/\alpha_3$ its indicial response is

$$\mu(p)\,\mathbf{1} = \mu_0 \frac{\alpha_1}{p + \alpha_1} \cdot \frac{\alpha_2}{p + \alpha_2} \cdot \frac{\alpha_3}{p + \alpha_3}\,\mathbf{1} \qquad (17)$$

Let the indicial response of the feedback path be

$$\beta_0(1 + Bp + Ap^2)\mathbf{1}.$$

then the indicial response of the amplifier with feedback is

$$h(p) = \frac{\mu_0 \alpha_1 \alpha_2 \alpha_3}{(p+\alpha_1)(p+\alpha_2)(p+\alpha_3) + \alpha_1\alpha_2\alpha_3\mu_0\beta_0(1+Bp+Ap^2)} \quad \ldots (18)$$

$$= \frac{\mu_0 \alpha_1 \alpha_2 \alpha_3}{(p+\lambda_1)(p+\lambda_2)(p+\lambda_3)} \quad \ldots \quad \ldots (19)$$

Where

$(p+\lambda_1), (p+\lambda_2), (p+\lambda_3)$

are the factors of the denominator of Equ. (18).

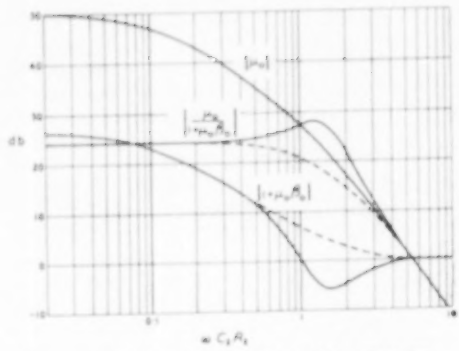


Fig. 11. *Amplitude characteristics of two-stage amplifier uncompensated (full line) and critically damped (dashed line).*

$$\therefore \lambda_1 + \lambda_2 + \lambda_3 = \alpha_1 + \alpha_2 + \alpha_3 + \alpha_1\alpha_2\alpha_3\mu_0\beta_0 A \quad \ldots (20a)$$

$$\lambda_1\lambda_2 + \lambda_2\lambda_3 + \lambda_3\lambda_1 = \alpha_1\alpha_2 + \alpha_2\alpha_3 + \alpha_3\alpha_1 + \alpha_1\alpha_2\alpha_3\mu_0\beta_0 B \quad (20b)$$

$$\lambda_1\lambda_2\lambda_3 = \alpha_1\alpha_2\alpha_3(1+\mu_0\beta_0) \quad \ldots \quad \ldots (20c)$$

The type of indicial response obtained depends on the nature of $\lambda_1$, $\lambda_2$ and $\lambda_3$: they are either all real, or two of them are conjugate complex numbers. If two of the factors are complex, the response will be oscillatory, so we need only consider the cases in which the factors are real. If $\lambda_1$, $\lambda_2$ and $\lambda_3$ are real and positive, equation (19) is of the same form as that for the gain of an amplifier without feedback having stages with time constants $1/\lambda_1$, $1/\lambda_2$, $1/\lambda_3$ (c.f. Equ. (17)). Since the time constants of the stages and the feedback factor determine the product $\lambda_1\lambda_2\lambda_3$ independently of the ratios $\lambda_1 : \lambda_2 : \lambda_3$ (Equ. 20c), it can be shown[8] that the smallest rise time is obtained for the indicial response when $\lambda_1 = \lambda_2 = \lambda_3$: this is the condition for the feedback amplifier to be critically damped.

5.1. *Three-stage amplifier with constant $\beta$*

If $\beta = \beta_0$ and $\lambda_1 = \lambda_2 = \lambda_3 = \alpha$, then

$$3\alpha = \alpha_1 + \alpha_2 + \alpha_3 \quad \ldots \quad \ldots \quad \ldots (21a)$$

$$3\alpha^2 = \alpha_1\alpha_2 + \alpha_1\alpha_3 + \alpha_2\alpha_1 \quad \ldots \quad \ldots (21b)$$

$$\alpha^3 = \alpha_1\alpha_2\alpha_3(1+\mu_0\beta_0) \quad \ldots \quad \ldots (21c)$$

These equations cannot be satisfied unless $\mu_0\beta_0 = 0$, or $\alpha_1 = \alpha_2 = \alpha_3 = 0$ or unless $\alpha_1$, $\alpha_2$ or $\alpha_3$ is complex. It is therefore impossible to obtain three equal factors to the denominator of the indicial response of a three-stage resistance-coupled amplifier with a negative-feedback path whose transfer coefficient is independent of frequency. Brockelsby and Mayr found that it was impossible to obtain a maximally-flat frequency response of the form $|A_0/A_{s0}|^2 = 1 + X^6$ (where $X$ is the normalized frequency) for a three-stage amplifier with constant $\beta$.

5.2. *Three-stage amplifier with a single time constant in the feedback path.*

If $\beta(p) = \beta_0 \frac{p+\alpha_4}{\alpha_4}$ and $\lambda_1 = \lambda_2 = \lambda_3 = \alpha$, then

$$3\alpha = \alpha_1 + \alpha_2 + \alpha_3 \quad \ldots \quad \ldots \quad \ldots (22a)$$

$$3\alpha^2 = \alpha_1\alpha_2 + \alpha_2\alpha_3 + \alpha_3\alpha_1 + \frac{\alpha_1\alpha_2\alpha_3}{\alpha_4}\mu_0\beta_0 \quad \ldots \quad \ldots (22b)$$

$$\alpha^3 = \alpha_1\alpha_2\alpha_3(1+\mu_0\beta_0) \quad \ldots \quad \ldots (22c)$$

$$\therefore \alpha = \tfrac{1}{3}(\alpha_1+\alpha_2+\alpha_3) = \sqrt[3]{\alpha_1\alpha_2\alpha_3(1+\mu_0\beta_0)} \quad \ldots \quad \ldots (23)$$

$$\alpha_4 = \frac{3\alpha_1\alpha_2\alpha_3\mu_0\beta_0}{\alpha_1^2+\alpha_2^2+\alpha_3^2-(\alpha_1\alpha_2+\alpha_2\alpha_3+\alpha_3\alpha_1)} \quad \ldots \quad \ldots (24)$$

$$1+\mu_0\beta_0 = \frac{(\alpha_1+\alpha_2+\alpha_3)^3}{27\alpha_1\alpha_2\alpha_3} \quad \ldots \quad \ldots (25)$$

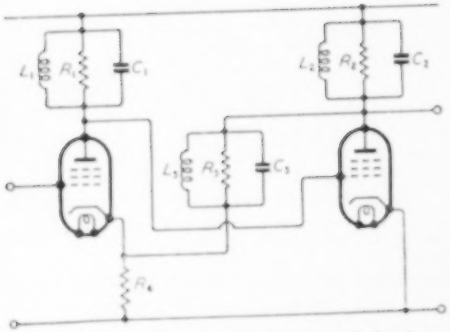


Fig. 12. *Band-pass circuit equivalent to the low-pass circuit of Figs. 2 and 5.*

It is seen that for any given time constants for the three stages of the amplifier, three equal factors can only be obtained for the particular value of the feedback factor given by Equ. (25).

5.3. *Three-stage amplifier with two time constants in the feedback path*

If $\beta(p)\mathbf{1} = \beta_0 \dfrac{p + \alpha_4}{\alpha_4} \dfrac{p + \alpha_5}{\alpha_5}\mathbf{1}$

and $\lambda_1 = \lambda_2 = \lambda_3 = \alpha$, then

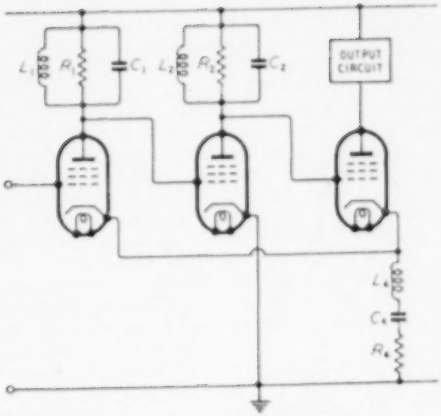


Fig. 13. *Band-pass circuit equivalent to the low-pass circuit of Figs. 3 and 6.*

$$3\alpha = \alpha_1 + \alpha_2 + \alpha_3 + \mu_0\beta_0 \frac{\alpha_1\alpha_2\alpha_3}{\alpha_4\alpha_5} \quad .. \quad (26a)$$

$$3\alpha^2 = \alpha_1\alpha_2 + \alpha_2\alpha_3 + \alpha_3\alpha_1 + \alpha_1\alpha_2\alpha_3 \frac{\alpha_4 + \alpha_5}{\alpha_4\alpha_5}\mu_0\beta_0 \quad .. \quad .. \quad (26b)$$

$$\alpha^3 = \alpha_1\alpha_2\alpha_3(1 + \mu_0\beta_0) \quad .. \quad .. \quad (26c)$$

$$\therefore \alpha_4, \alpha_5 = \frac{1}{2a}\left(b \mp \sqrt{b^2 - 4ac}\right) \quad .. \quad (27)$$

where

$$a = 3\alpha - (\alpha_1 + \alpha_2 + \alpha_3)$$

$$b = 3\alpha^2 - (\alpha_1\alpha_2 + \alpha_2\alpha_3 + \alpha_3\alpha_1)$$

$$c = \alpha^3 - \alpha_1\alpha_2\alpha_3$$

Three equal factors can therefore be obtained for any given time constants for the three stages provided that

$$1 + \mu_0\beta_0 > \frac{(\alpha_1 + \alpha_2 + \alpha_3)^3}{27\alpha_1\alpha_2\alpha_3}$$

When

$$1 + \mu_0\beta_0 = \frac{(\alpha_1 + \alpha_2 + \alpha_3)^3}{27\alpha_1\alpha_2\alpha_3},$$

$\alpha_4$ or $\alpha_5$ becomes infinite and the circuit reduces to that of Sec. 5.2 having only one time constant in the feedback path.

$$\text{When } 1 + \mu_0\beta_0 < \frac{(\alpha_1 + \alpha_2 + \alpha_3)^3}{27\alpha_1\alpha_2\alpha_3},$$

equal factors cannot be obtained and the transient response is over damped and less rapid.

When equation (27) is satisfied

$$h(p)\mathbf{1} = \frac{\mu_0\alpha_1\alpha_2\alpha_3}{(p + \alpha)^3}\mathbf{1}$$

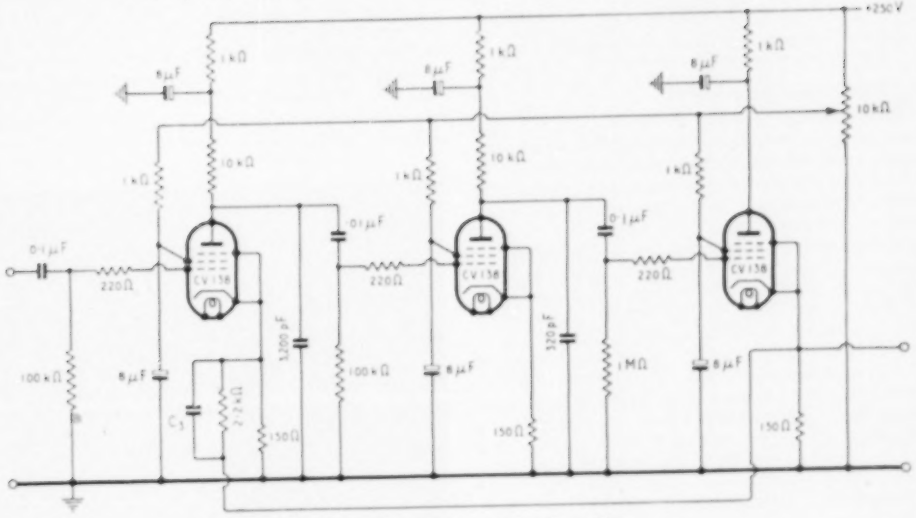


Fig. 14. *Circuit of experimental amplifier.*

where

$$\alpha^3 = \alpha_1\alpha_2\alpha_3(1 + \mu_0\beta_0) \quad .. \quad .. \quad (26c)$$

$$\therefore h(p) = \frac{\mu_0}{1 + \mu_0\beta_0}\cdot\frac{\alpha^3}{(p + \alpha)^3} \quad .. \quad (15)$$

and

$$h(t) = \frac{\mu_0}{1 + \mu_0\beta_0}\left[1 - e^{-\alpha t}\left(1 + \alpha t + \frac{\alpha^2 t^2}{2}\right)\right], t > 0 \quad .. \quad .. \quad .. \quad (16)$$

A particularly simple special case is obtained when

$$\alpha_2 = \alpha_3 = \alpha_1(1 + \mu_0\beta_0),$$

then $\alpha = \alpha_4 = \alpha_5 = \alpha_6 = \alpha_7 = \alpha_1(1 + \mu_0\beta_0)$.

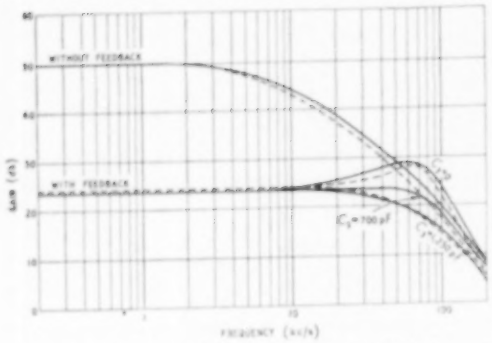


Fig. 15. *Measured (solid line) and calculated (dashed line) response curves for the amplifier of Fig. 14.*

## 6. Band-pass Amplifiers

The amplifier circuits so far considered have had low-pass frequency characteristics. A low-pass circuit can be transformed to an equivalent band-pass circuit by replacing each inductance with a series resonant circuit and each capacitance with a parallel resonant circuit[7].

The envelope of the response of the band-pass circuit to an input voltage of the mid-band frequency, amplitude-modulated by a unit step, is then approximately the indicial response of the equivalent low-pass circuit[7].

The band-pass equivalent of the circuit of Fig. 2, modified as in Fig. 5, is shown in Fig. 12, each resonant circuit being tuned to the same frequency. The envelope of its response is therefore given by equation (11) if the time constant of the feedback path is given by equation (13). For the band-pass circuit, the time constants of the stages are given by $1/\alpha_1 = 2R_1C_1$, etc.

The band-pass equivalent of the circuit of Fig. 3, modified as in Fig. 6, is shown in Fig. 13, each resonant circuit being tuned to the same frequency. This circuit has recently been described by Clifford[8], and is claimed to be remarkable for its simplicity and high performance.

## 7. Experimental Results

Fig. 14 shows an amplifier which was built in order to obtain experimental verification of the theory. Capacitors were connected between the anode of each stage and earth so that the total interstage capacitances (including valve and wiring capacitances) had the values shown. The time constant of one stage was ten times that of the other. The screen voltage of the valves was adjusted so that the gain at low frequencies with the feedback path disconnected was 50 db, then $\mu_0\beta_0 = 20$ and Fig. 11 should show the gain-frequency characteristics of the amplifier with and without feedback. The value of capacitance required to be connected across the feedback resistor in order to obtain critical damping [calculated from equation (13)] was 1250 pF.

Fig. 15 shows the measured gain-frequency response of the amplifier without feedback, with an uncompensated feedback path and with capacitance connected across the feedback resistor. Fair agreement is obtained between the measured and calculated values.

Fig. 16 shows oscillograms of the response of the amplifier to a 35-$\mu$s rectangular pulse. The response of the uncompensated amplifier is oscillatory. When the feedback resistor is shunted by a 700-pF capacitor, the response is still oscillatory but the overshoot is less and the oscillation decays more rapidly. When the feedback resistor is shunted by 1250-pF, the response is critically damped. When the feedback resistor is shunted

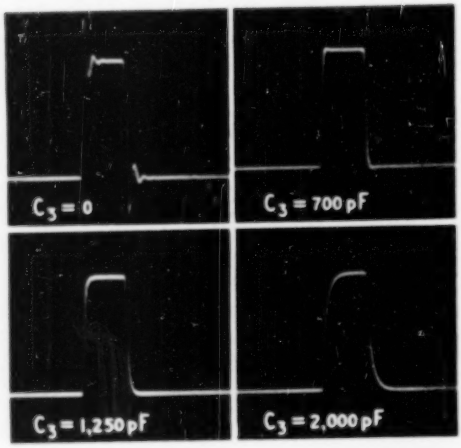


Fig. 16. *Response to a 35-$\mu$sec pulse of the amplifier of Fig. 14.*

by 2000-pF, the response is over-damped and less rapid.

## 8. Acknowledgment

Acknowledgment is made to the Engineer-in-Chief of the G.P.O. and to the Controller of H.M. Stationery Office for permission to publish this paper.

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# INPUT IMPEDANCE OF TWO CROSSED DIPOLES

By LL. G. Chambers, M.Sc., A.Inst.P.

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**SUMMARY.**—An application of Carter's Method is made to the calculation of the mutual impedance of a pair of crossed dipoles. This is used to give the input impedance of such a crossed pair. It is found that, to a good degree of approximation, the mutual (and input) impedances may be expressed as very simple functions of the angle of separation of the dipoles.

## 1. Introduction

IN recent years several papers on the calculation of the input impedance of various aerials have appeared. Up to the present, however, calculations on the mutual impedance of a pair of crossed dipoles (see Fig. 1) do not appear to have been made, and it is the purpose of this note to fill this gap in aerial theory. The calculations have been carried through for a pair of infinitely-thin dipoles, each having a total length of half a wavelength. It is thought that the *mutual* impedance would not be very sensitive to the thickness of the wire forming the dipoles.

Fig. 1. *Pair of crossed dipoles.*

## 2. Field of a Single Dipole

If we have a straight wire, extending along the $z$ axis between $z = a$ and $z = b$, and the charge $\sigma$ per unit length and the current $I$ both satisfy the wave equation

$$\frac{\partial^2 f}{\partial z_0^2} = \frac{1}{c^2}\frac{\partial^2 f}{\partial t_0^2} \quad .. \quad .. \quad .. \quad .. \quad (1)$$

($z_0$, $t_0$ being current co-ordinates along the wire), then it has been shown by A. G. D. Watson[1] that

$$4\pi\epsilon E_z = \left[\frac{\sigma\left(z_0, t - \frac{r'}{c}\right)}{r'}\right]_{z_0=a}^{z_0=b} \quad .. \quad .. \quad (2)$$

$$4\pi\epsilon E_\rho = \left[\frac{\sigma\left(z_0, t - \frac{r'}{c}\right)z_0 - z}{r'\rho}\right]_{z_0=a}^{z_0=b} \quad .. \quad (3)$$

where $\epsilon$ is the dielectric constant of the medium in which the wire is immersed. The other symbols are explained by Fig. 2.

Simple vector resolution gives us the result that

$$E_r = E_\rho \cos\theta + E_z \sin\theta \quad .. \quad .. \quad (4)$$

We now consider the application of this theory to a dipole (Fig. 3). Let us suppose the current distribution in the dipole is given by

$$I = I_0 e^{j\omega t}\frac{\sin k_0(l - |z_0|)}{\sin k_0 l}, \left(k_0 = \frac{\omega}{c}\right) \quad .. \quad (5)$$

Now $I$ and $\sigma$ are related by

$$\frac{\partial\sigma}{\partial t_0} + \frac{\partial I}{\partial z_0} = 0 \quad (6)$$

$$\frac{\partial\sigma}{\partial z_0} + \frac{1}{c^2}\frac{\partial I}{\partial t_0} = 0 \quad (7)$$

Equation (6) is the equation of continu-

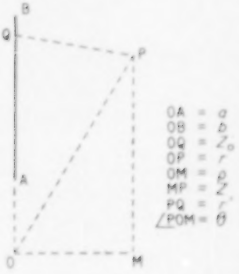


Fig. 2. *Co-ordinates for a straight wire; OB is the z-axis.*

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ity, and (7) is the condition that $I$ and $\sigma$ satisfy (1). We note that

$$c = 1/\sqrt{(\mu\epsilon)} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

and we write

$$R_s = \frac{1}{4\pi}\sqrt{\frac{\mu}{\epsilon}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

For free space,

$$R_s \approx 30 \text{ ohms} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

Referring to Fig. 3, it is obvious that

$$r_-^2 = r^2 - 2rl\sin\theta + l^2 \quad \ldots \quad \ldots \quad (11)$$

$$r_+^2 = r^2 + 2rl\sin\theta + l^2 \quad \ldots \quad \ldots \quad (12)$$

At P

$$4\pi\epsilon E_r = \left[\frac{\sigma(l)}{r_-}e^{-jk_0r_-} - \frac{\{\sigma(+0) - \sigma(-0)\}e^{-jk_0r}}{r} - \frac{\sigma(-l)e^{-jk_0r_+}}{r}\right] \quad \ldots \quad (13)$$

and $4\pi\epsilon E_\theta =$

$$\left[\frac{\sigma(l)}{r_-}e^{-jk_0r_-}\cdot\left(\frac{l - r\sin\theta}{r\cos\theta}\right) - \frac{\sigma(-l)}{r_+}e^{-jk_0r_+}\left(\frac{-l - r\sin\theta}{r\cos\theta}\right) - \frac{\{\sigma(+0) - \sigma(-0)\}e^{-jk_0r}}{r}\left(\frac{-r\sin\theta}{r\cos\theta}\right)\right] \quad (14)$$

where we have now dropped the time factor from (5) and (6).

We deduce

$$\sigma(l) = -\sigma(-l) = \frac{jI_0}{c} \quad \ldots \quad \ldots \quad (15)$$

$$\sigma(+0) = -\sigma(-0) = -\frac{jI_0}{c}\cos k_0 l \quad \ldots \quad (16)$$

The self-impedance $Z_s$ is given by

$$Z_s = -\frac{1}{4\pi\epsilon}\cdot\frac{1}{\sin^2 k_0 l}\int_{-l}^{l}-\frac{j}{c}\left[\frac{e^{-jk_0(l-r)}}{l-r} + \frac{e^{-jk_0(l+r)}}{l+r} + 2\cos k_0 l\,\frac{e^{-jk_0r}}{r}\right]\times \sin k_0(l - |r|)\,dr \quad \ldots \quad \ldots \quad (17)$$

$$= \frac{jR_s}{\sin^2 k_0 l}\int_{-l}^{l}\left\{\frac{e^{-jk_0(l-r)}}{l-r} + \frac{e^{-jk_0(l+r)}}{l+r}\right\}\sin k_0(l - |r|)\,dr + 2R_s\frac{\cos k_0 l}{\sin^2 k_0 l}\int_{-l}^{l}\frac{\sin k_0 r}{r}\sin k_0(l - |r|)\,dr \quad \ldots \quad \ldots \quad \ldots \quad (18)$$

If we suppose $k_0 l = \pi/2 + \alpha\pi$, $\alpha$ being integral, $Z_s$ may be shown to reduce to

$$R_s\int_0^{\infty}\frac{1 - e^{-jx}}{x}\,dx$$

$$= R_s\{(\gamma + \log 4k_0 l - \mathrm{Ci}\,4k_0 l) + j\,\mathrm{Si}\,4k_0 l\} \quad (19)$$

which is a well known result (2).

Now if we assume the dipole to be at any resonance the current

$$= I_0\frac{\sin k_0(l - |r|)}{\sin k_0 l} = I_0\cos k_0 r \quad (20)$$

no matter what resonance is occurring.

The field component $E_r$ is given by

$$E_r = E_\theta\cos\theta + E_z\sin\theta$$

Fig. 3. *Co-ordinates of dipole.*

$$= \frac{1}{4\pi\epsilon}\left[\frac{\sigma(l)}{r_-}e^{-jk_0r_-}\cdot\frac{l}{r} + \frac{\sigma(l)}{r_-}e^{-jk_0r_-}\cdot\frac{l}{r}\right] \quad (21)$$

## 3. Calculation of Mutual Impedance

The mutual impedance between two dipoles placed at an angle $\pi/2 - \theta$ to one another is given by integrating the product of the field due to one, tangential to the second, and the current distribution along the second. This process gives us

$$Z_m = -\frac{1}{4\pi\epsilon}\cdot\frac{1}{\sin^2 k_0 l}\cdot\frac{-j}{c}\times \int_{-l}^{l}\left(\frac{e^{-jk_0r_+}}{r_+} - \frac{e^{-jk_0r_-}}{r_-}\right)\frac{l}{r}\sin k_0(l - |r|)\,dr \quad (22)$$

provided the current distribution in both dipoles is given by (5). If the dipoles are both in resonance we obtain

$$\frac{Z_m}{R_s} = 2j\int_0^{l}\frac{1}{R}\left\{\frac{e^{-jR_+\pi/2}}{R_+} - \frac{e^{-jR_-\pi/2}}{R_-}\right\}\cos\pi/2\,R/l\,dR = \alpha + j\beta \quad \ldots \quad (23)$$

where $r = Rl$, $r_+ = R_+ l$, $r_- = R_- l$

While it is impossible to integrate exactly, these integrals may easily be evaluated by quadrature. The results are given in Table 1 for the argument $\phi = \pi/2 - \theta$ and plotted in Fig. 4.

It is obvious that

$$Z_m\left(\frac{\pi}{2} + \theta\right) = -Z_m\left(\frac{\pi}{2} - \theta\right) \quad \ldots \quad (24)$$

It will be seen that to within 1.5 per cent.

$$\alpha = 1.219\cos\phi \quad \ldots \quad \ldots \quad \ldots \quad (25)$$

$$\beta = 2.556 - 0.0284\phi \quad \ldots \quad \ldots \quad \ldots \quad (26)$$

Where, in (26), $\phi$ is measured in degrees.

It may be shown that, for $kd = \pi/2$

$$Z_m = 2R_s\{1.219 + 0.709j\} \quad .. \quad .. \quad (27)$$

TABLE 1

$Z_m = 2R_s(\alpha + j\beta)$

| $\phi$ | $\alpha$ | $\beta$ |
|---|---|---|
| 0° | 1.219 | 2.361 |
| 15° | 1.160 | 2.127 |
| 30° | 1.054 | 1.704 |
| 45° | 0.859 | 1.269 |
| 60° | 0.607 | 0.841 |
| 75° | 0.314 | 0.418 |
| 90° | 0 | 0 |

## 4. Calculation of Input Impedance

If we have a current $I$ divided between two branches of a system, then the equations satisfied are

$$V = I_1 Z_{11} + I_2 Z_{21}$$
$$= I_1 Z_{12} + I_2 Z_{22} \quad .. \quad .. \quad .. \quad (28)$$

The impedance of this two-branch system is given by

$$V = ZI \quad .. \quad .. \quad .. \quad .. \quad (29)$$

and, if the network is symmetrical,

$$Z_{11} = Z_{22} = Z_s; Z_{21} = Z_{12} = Z_m;$$
$$I_1 = I_2 = \tfrac{1}{2} I \quad .. \quad .. \quad .. \quad (30)$$

giving

$$2Z = Z_s + Z_m \quad .. \quad .. \quad .. \quad (31)$$

Applying this to our dipole system we have

$$\frac{Z}{R_s} = 1.219\,(1 + \cos\phi) + j(3.265 - 0.0284\phi^\circ)$$
$$.. \quad .. \quad .. \quad (32)$$

using our approximate relation for $Z_m$. It will be seen that the radiation resistance vanishes for $\phi = \pi$ (as might have been expected) and that the imaginary part vanishes for $\phi = 115°$. It should be noted that these results depend on the current distribution in the dipoles being given by (5).

It is thought that the mutual impedance between two such dipoles should not vary appreciably with their thickness, provided that it is small compared with the length. These results could be applied to a 'fan' of dipoles in a similar manner, provided that the ratio of the maximum currents in each dipole is assumed known.

It will be noticed that the imaginary part of the impedance of a single dipole differs from that of two dipoles separated by a small angle. This is due to the fact that fields may exist

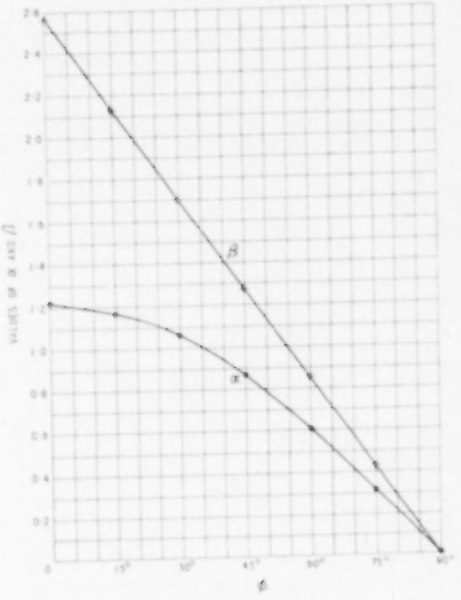


Fig. 4. *Plots of the expressions* $\alpha = 1.219 \cos\phi$ *and* $\beta = 2.356 - 0.0284\ \phi$.

between the dipoles which, while they affect the reactance, do not affect the resistive part of the impedance which depends only on the field at infinity. This is substantially the same for a single dipole and a pair of dipoles at a very small angle.

## 5. Acknowledgment

I am indebted to the Admiralty for permission to publish this note.

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# DESIGN OF CATHODE-COUPLED AMPLIFIERS

By S. G. F. Ross, B.E.

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**SUMMARY.**—This paper contains a brief introductory outline of the advantages and possibilities of cathode-coupled amplifiers. Some recent papers are reviewed, and certain short-comings pointed out.

A rigorous theoretical analysis is established, and some experimental results given to show that this analysis can be relied on within the accuracy of published valve data.

## 1. Introduction

THE use of the normal cathode-follower as a wideband amplifier is well established. Its high input impedance and low output impedance render it capable of good wideband performance, at the same time providing an excellent transformer action where it is necessary to match a high-impedance load into a low-impedance output cable, as in television work.

On the other hand there is a great demand in many fields, such as television, for low noise-level amplifiers. Although triodes are superior in this respect to pentodes, they suffer from lack of sufficient gain, and are subject to adverse Miller effect. The latter difficulty can be overcome by the use of the earthed-grid, or inverted triode. This has found some use in push-pull power-amplifier stages, and is also useful for matching low-impedance input cables into high-impedance receiver circuits, by virtue of its low input impedance.

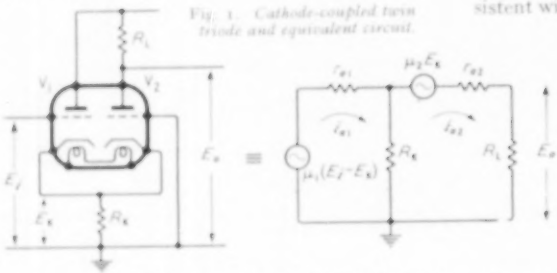


*Fig. 1. Cathode-coupled twin triode and equivalent circuit.*

The nature of the impedance transformations of the cathode-follower and inverted amplifier opens up the possibility of combining these two amplifiers with a common cathode load. This is what is done in the cathode-coupled (or 'long-tailed') amplifier. Such a combination, using a single twin-triode amplifier, is found to give wideband amplification comparable with that of a pentode, but with very superior signal-to-noise ratio.

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Several such amplifier stages may be used in cascade in television-receiver circuits. Since the same valve types are applicable to synchronizing and other circuits, this makes for a marked saving in the number of valve types employed, as well as simplifying the coupling circuits required in the video amplifier or r.f. stages employing these valves.

The increasing popularity of this form of amplifier makes it desirable to derive a theoretical analysis which gives a close approximation to experimental results.

Several previous attempts to do this have failed because of the assumption of constant valve parameters. Sziklai and Schroeder[1] made a good start on the problem of the cathode-coupled twin triode. However, careful analysis of the equivalent circuit and valve characteristics shows that their assumption that the valve parameters are equal for each triode section is not justified. For the same reason, Pullen[2] failed to obtain a set of theoretical curves consistent with those he had obtained experimentally.

## 2. Graphical Analysis

It is as well to mention here a fairly simple and straightforward graphical analysis evolved by Rifkin.[3] This makes use of dynamic valve characteristics which are, however, difficult to determine accurately. The difficulty here arises from the fact that the combined bias voltage lowers the available anode-cathode potential by an equal amount. To take this into account for any combination of anode load and cathode resistor requires a multiplicity of dynamic curves. Neglect of this effect can introduce substantial errors, especially at low values of anode current, where the values of $r_a$ and $\mu$ change rapidly with the current.

## 3. Theoretical Design Data

The following method of approach, although rather more tedious than that of Rifkin, nevertheless lends itself to easy tabulation and curve

production. At the same time it is strictly theoretical and consequently is preferable to the more approximate graphical method. Referring to Fig. 1, the formula for the voltage amplification of the cathode-coupled amplifier (as verified in the Appendix) is :

$$A = \frac{E_0}{E_i} = \frac{\mu_1 (\mu_2 + 1) R_L R_K}{r_{a1} r_{a2} + r_{a1} R_L + r_{a2} R_K (\mu_2 + 1) + R_K (\mu_1 + 1) (r_{a2} + R_L)}$$

Application of this formula is dependent on the determination of $i_{a1}$ and $i_{a2}$ for any chosen values of $R_L$ and $R_K$, thereby obtaining $r_{a1}$, $r_{a2}$, $\mu_1$ and $\mu_2$ from the valve curves (Fig. 3). The procedure is quite straightforward and is indicated here for a 6SN7 valve using the recommended anode supply of 250 volts.

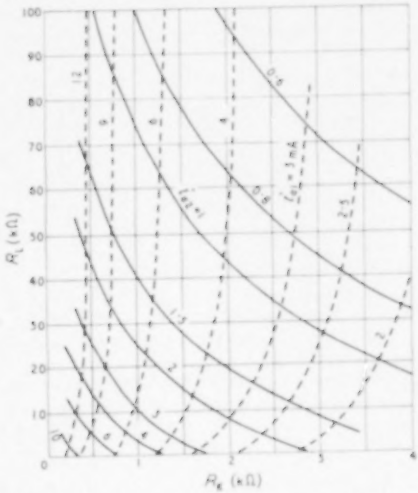



Fig. 2. *Anode-current curves for cathode-coupled 6SN7.*

Select $i_{a2} = 3$mA (constant).
Let bias $= -8$ V.
Then anode-cathode potential of $V_1 = 242$ V.
Whence, from published curves, $i_{a1} = 8$ mA.

$$\therefore R_K = \frac{8}{3 + 8} \times 1{,}000 = 727\ \Omega.$$

Again, from published curves, anode-cathode potential of $V_2 = 194$ V.

$$\therefore \text{Voltage drop in } R_L = (242 - 194) = 48 \text{ V.}$$

$$\therefore R_L = \frac{48}{3} \times 1000 = 16{,}000\ \Omega$$

Table 1 shows in skeleton form how a series of these results can be easily tabulated :

Using this Table, we can draw a set of $R_L - R_K$ curves for constant values of $i_{a2}$. Since the $i_{a1}$ column is identical for each chosen value of $i_{a2}$, there will be corresponding $i_{a1}$ points on each curve plotted and, by simple interpolation, a further set of curves may be plotted of constant values of $i_{a1}$. This process is indicated in Fig. 2, which shows clearly where it is necessary to choose $i_{a2}$ values very closely together in forming the Table.

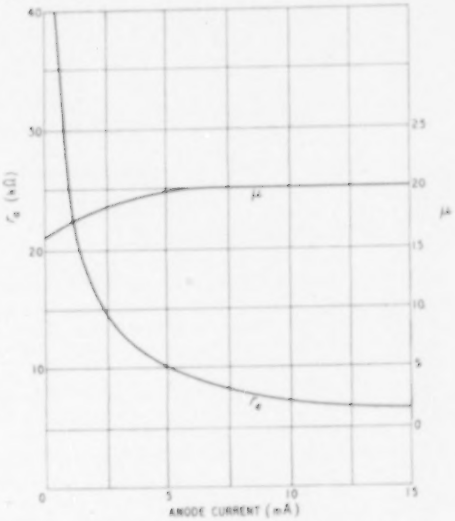



Fig. 3. *$r_a$ and μ characteristics of 6SN7.*

TABLE 1

| BIAS (− V) | $i_{a1}$ (mA) | $i_{a2}$ (mA) | $R_L$ (Ω) | $R_K$ (Ω) | $i_{a2}$ | $R_L$ | $R_K$ | $i_{a2}$ | $R_L$ | $R_K$ | $i_{a2}$ | $R_L$ | $R_K$ |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 6.0 | 12.8 | 0.8 | 151,800 | 440 | 1.0 | 116,700 | 435 | | | | 6.0 | 9,800 | 320 |
| 6.5 | 11.6 | | 140,300 | 520 | | 107,100 | 520 | | | | | 8,100 | 370 |
| 7.0 | 10.4 | | 129,400 | 625 | | 97,500 | 610 | | | | | 6,300 | 425 |
| 7.25 | 9.8 | | 124,000 | 680 | | 92,800 | 670 | | | | | 5,500 | 460 |
| 11.0 | 2.1 | | 34,350 | 3,750 | | 20,800 | 3,550 | | | | — | — | — |

When this set of curves has been established it becomes a simple matter to compile a table, such as shown in skeleton form in Table 2, giving $r_{a1}$, $r_{a2}$, $\mu_1$ and $\mu_2$ and thence the voltage amplification for any chosen values of $R_L$ and $R_k$. From this Table we can draw up a final set of curves (the full lines in Fig. 4) from which to choose suitable circuit constants for any required gain.

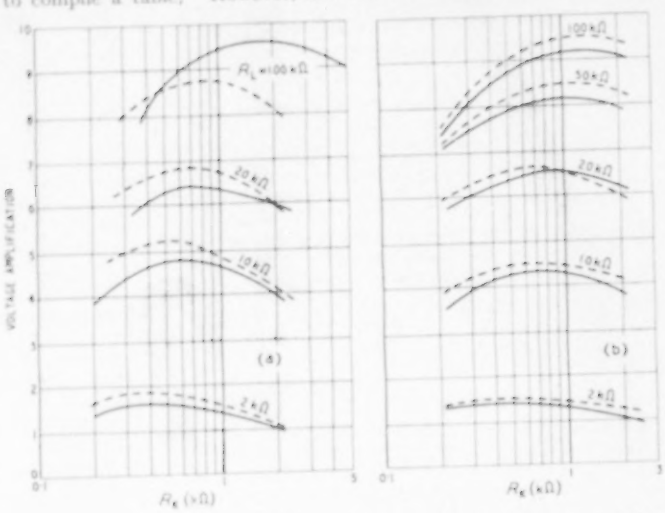



Fig. 4. (a) *Comparison of theoretical (full curves) and experimental (broken curves) amplification for cathode-coupled 6SN7.* (b) *comparison of experimental amplification and theoretical curves (full lines) obtained by plotting individual valve characteristics.*

## 4. Experimental Verification

The curves of Fig. 4 indicate that the theoretical gain is somewhat higher than that obtained experimentally by Pullen. However, he has not stated the anode supply voltage, nor the magnitude of the input signal used. The author has carried out some tests with a single 6SN7, using the same anode supply (regulated 250 volts) as that used in the theoretical treatment, and an input signal of 0.5 volt. The curves obtained are shown by the broken lines in Fig. 4(a). Further tests showed that the amplification dropped off as the magnitude of the input signal was increased, as indicated in Fig. 5. This is due to the unbalanced effect of $r_a$ on the positive and negative half-cycles of current.

The above tests were carried out at a frequency of 100 kc/s, and this accounts for the departure from normal with high values of anode load. However, the results show that, within the range normally applicable to wideband amplifiers, the agreement between experimental and theoretical results is very good.

## 5. Precise Test of Accuracy of Theory

The author found that there were appreciable variations of characteristics from valve to valve, in the low cathode-current zones outside the normal operating range. In addition, the static value of $r_a$ is a function, not only of anode current, but also of anode voltage and grid bias. In view of this, and the fairly high frequency employed in the previous test, it was considered that a more exhaustive test of the accuracy of the new formula was desirable.

A sample 6SN7 twin triode was taken, its characteristic curves plotted, and from these the curves of constant $i_a$ derived (as in Fig. 2).

TABLE 2

| $R_L$ | $R_k$ | $i_{a1}$ | $\mu_1$ | $r_{a1}$ | $i_{a2}$ | $\mu_2$ | $r_{a2}$ | $A$ |
|---|---|---|---|---|---|---|---|---|
| 2,000 | 200 | 14.0 | 20.0 | 6,300 | 12.0 | 20.0 | 6,700 | 1.40 |
| | 2,000 | 2.75 | 18.9 | 13,300 | 2.5 | 18.7 | 14,000 | 1.10 |
| 10,000 | 200 | 15.0 | 20.0 | 6,450 | 7.2 | 20.0 | 8,300 | 3.80 |
| | 2,000 | 3.2 | 19.2 | 12,300 | 1.9 | 18.3 | 16,000 | 4.02 |
| 20,000 | 400 | 12.0 | 20.0 | 6,700 | 3.65 | 19.3 | 11,550 | 6.13 |
| etc. | 2,000 | 3.3 | 19.3 | 11,800 | 1.5 | 17.9 | 18,400 | 5.94 |

Having thus obtained values of $i_{s1}$ and $i_{s2}$ for the chosen values of $R_L$ and $R_K$, the valve parameters for these values were obtained from the characteristic curves for the correct operating points (as regards bias or anode voltage).

These new values were then employed in the formula to obtain the theoretical gain curves. These are plotted in Fig. 4(b) together with experimental curves, obtained by using an input signal of 0.5 volt at a frequency of 1,000 cycles per second.

Allowing for small instrumental errors (these were minimized by the use of the lower frequency) and errors in plotting the characteristics and deriving their slopes, it would seem the agreement of the experimental curves to within 5 per cent of all theoretical gain curves is quite conclusive proof of the first-order accuracy of the formula.

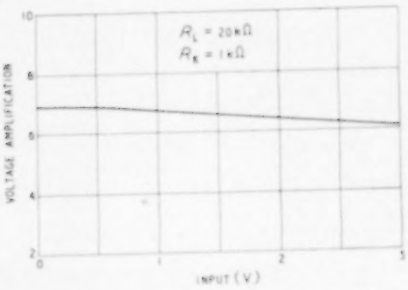



Fig. 5. *Amplification as a function of input signal.*

## 6. Conclusions

The foregoing precise analysis leaves us with the assurance that the new formula is accurate. In addition, the curves of Fig. 4 show that the experimental curves can be reasonably duplicated theoretically by applying the formula to published valve data. Consequently, we are free to employ this method, which is as accurate as required for design purposes, and permits interchangeability of valves.

## 7. Acknowledgment

This work was carried out at the University of Adelaide, under a grant from the Council for Scientific and Industrial Research of Australia, and under the direction of Professor E. O. Willoughby. All assistance given is gratefully acknowledged.

### APPENDIX

Verification of the formula used in this paper is a matter of simple algebra.

Considering the equivalent circuit of Fig. 1 —

$$E_K = (i_{s1} - i_{s2})\, R_K.$$

$$\mu_1(E_i - E_K) - i_{s1}\, r_{s1} - (i_{s1} - i_{s2})\, R_K = 0$$

or $\mu_1 R_K(i_{s1} - i_{s2}) + i_{s1}\, r_{s1} + (i_{s1} - i_{s2})\, R_K = \mu_1 E_i$

$$\therefore i_{s1}[R_K(\mu_1 + 1) + r_{s1}] - i_{s2} R_K(\mu_1 + 1) = \mu_1 E_i$$

$$\text{or } i_{s1} - i_{s2}\cdot\frac{R_K(\mu_1+1)}{R_K(\mu_1+1)+r_{s1}} = \frac{\mu_1 E_i}{R_K(\mu_1+1)+r_{s1}} \qquad (1)$$

Again,

$$\mu_2 E_K - i_{s2}(r_{s2}+R_L) - (i_{s2} - i_{s1})\, R_K = 0$$

$$\text{or } \mu_2 R_K(i_{s1} - i_{s2}) - i_{s2}(r_{s2}+R_L) - (i_{s2} - i_{s1})\, R_K = 0$$

$$\therefore i_{s1} R_K(\mu_2+1) - i_{s2}[[R_K(\mu_2+1) + r_{s2} + R_L]] = 0$$

$$\text{or } i_{s1} - i_{s2}\cdot\frac{R_K(\mu_2+1)+r_{s2}+R_L}{R_K(\mu_2+1)} = 0 \qquad (2)$$

Subtracting (2) from (1)

$$i_{s2}\left\{\frac{(R_K(\mu_2+1)+r_{s2}+R_L}{R_K(\mu_2+1)} - \frac{R_K(\mu_1+1)}{R_K(\mu_1+1)+r_{s1}}\right\} = \frac{\mu_1 E_i}{R_K(\mu_1+1)+r_{s1}}$$

$$\therefore i_{s2} = \frac{\mu_1 E_i R_K(\mu_2+1)}{r_{s1} r_{s2} + r_{s1} R_L + r_{s1} R_K(\mu_2+1) + R_K(\mu_1+1)(r_{s2}+R_L)}$$

$$\therefore A = \frac{E_0}{E_i} = \frac{i_{s2} R_L}{E_i}$$

$$= \frac{\mu_1(\mu_2+1) R_L R_K}{r_{s1} r_{s2} + r_{s1} R_L + r_{s1} R_K(\mu_2+1) + R_K(\mu_1+1)(r_{s2}+R_L)}$$

### REFERENCES

1 = Cathode-coupled Wide-band Amplifiers," by G. C. Sziklai and A. C. Schroeder, *Proc. Inst. Radio Engrs*, Vol. 33, Oct. 1945, pp. 701–709.

2 = The Cathode-coupled Amplifier," by K. A. Pullen, Jr, *Proc. Inst. Radio Engrs*, Vol. 34, pp. 402–407.

3 = A Graphical Analysis of the Cathode-Coupled Amplifier," by M. S. Rifkin, *Communications*, Dec. 1946, pp. 16, 42.

## INSTITUTE OF PHYSICS

The Summer Meeting of the Industrial Radiology Group will be held at the New Horticultural Hall, Victoria, London, S.W.1., from 25th–28th July 1950. There will also be a Technical Exhibition of the International Congress of Radiology and admission to both will be by ticket, obtainable free from the Institute of Physics, 47, Belgrave Sq., London, S.W.1.

## 1950 S.I.M.A. EXHIBITION AND SYMPOSIUM

The Electronics section is holding its Annual Exhibition and Symposium at the Examination Hall, Queen Square, London, W.C.1., from Tuesday 5th September to Friday 8th September.

The following papers will be read :

5th (afternoon), " Electronic Optics and the Use of Materials," by Professor G. I. Finch.

6th (morning), " Recent Improvements in Direct Recording," by W. Bamford.

6th (afternoon), " The Vibrating Reed Electrometer and its Application to the Measurement of Small Ionization Currents," by R. Y. Parry and H. W. Finch ; " Recent Improvements in pH Measurement," by Electronic Instruments Ltd.

7th (morning), " Industrial Batching and Counting," by Cinema Television Ltd.

7th (afternoon), " The Measurement of Thickness by Means of Beta Rays, " by Baldwin Instrument Co., Ltd.

8th (morning), " User Problems in Electronic Instrumentation," by Imperial Chemical Industries Ltd.

Entrance to the Symposium and Exhibition will be by ticket obtainable on application to the Secretary of the Scientific Instrument Manufacturers' Association, 17 Princes Gate, London, S.W.7. Entrance to the exhibition only can be secured on presentation of a trade card.

# CORRESPONDENCE

*Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.*

## Can $H$ and $D$ be Measured Directly?

SIR,—In the April Editorial you have criticized our methods of measuring $H$ and $D$, described in *Philips' Research Reports*, April 1949. In this criticism a distinction was made between the directly measurable magnitudes $B$ and $E$, and the only calculable magnitudes $H$ and $D$. This distinction is elucidated in *Wireless Engineer*, August 1946, p. 207 as follows: "$B$ and $E$ can be determined by measurements at the point"; $D$ and $H$ "have to be calculated on certain assumptions as to the distribution of . . . electric charge over an area or of the ampere-turns over the magnetic path-length." In *Nature*, January 9th, 1937, p. 46, the same idea is formulated in the following way: "It will be seen from the above formulae that $E$ and $B$ are measurable characteristics of the electric and magnetic fields, the former by the force on a stationary charge and the latter by the force on a current or moving charge irrespective of the medium. $D$ and $H$, however, are not directly measurable concepts . . . $D$ is calculable at any point by dividing a charge . . . by an area; and similarly the magnetizing force is calculable by dividing a current by a length."

In our opinion, only the following statement is correct: It is possible to measure $E$ and $B$ "directly" with the aid of forces, whereas a "direct" measuring of $D$ and $H$ cannot be performed by forces. The distinction described in the quoted passages, however, is not real, as we hope to make clear in the following.

The definition of $E$, on which the usual measurement is based, reads:

$$E = \lim_{Q} \frac{F}{Q},$$

for the spatial dimensions of $Q$ approaching zero. ($F$ = force; $Q$ = charge).

The definition of $B$ is:

$$B = \lim_{I\Delta s \to 0} \frac{F}{I\Delta s}$$

($I\Delta s$ = element of current).

It is evident that even $E$ and $B$ "are not directly measurable concepts, but calculable by dividing a force by a charge, or by a current element or moving charge." Thus in this respect there is no essential difference between our methods of measuring $D$ and $H$, and the conventional methods of measuring $E$ and $B$.

According to your Editorial, the transition to infinitesimally small areas and lengths in the processes of measuring $D$ and $H$ is considered as "a very fictitious experiment." But a similar transition is necessary for measuring $E$ and $B$: In measuring $E$ the charge $Q$ has to be concentrated at a point; in measuring $B$ the current-element must be infinitesimally short and thin. In this respect there is thus no essential difference between the four methods of measuring either.

Yet a slight distinction might be made. In the concepts of either $D$ or $H$, the transition to infinitesimally small magnitudes is essential for the concept, whereas in the concepts of $E$ and $B$ this transition may be considered as an important, but not essential, condition. This reasoning can be justified in the following manner. If there were charges and current-elements of infinitesimally small dimensions, it would be possible to measure $E$ exactly by a charge of finite value and also $B$ by a current element of finite value. The values of $D$ and $H$, on the other hand, can be measured only approximately with discs or solenoids of finite dimensions. This difference, however, seems to us a very subtle one. Therefore we conclude that in vacuo $D$ and $H$ can be measured as well as $E$ and $B$. In our paper, we wished only to deny the correctness of the conclusions drawn from the (false) statement that $D$ and $H$ in vacuo are not measurable as separate magnitudes.

As to your remarks concerning the measuring of $D$ and $H$ in matter, we are glad to state that there is no disagreement between you and us. In media without remanence and hysteresis, these methods are good; in a medium with residual magnetism the ideal solenoid must be filled with an unmagnetic insulator, so that we have to do with a cavity experiment, as you already suggested for our $D$-measuring.

P. CORNELIUS,
H. C. HAMAKER.

Philips' Research Laboratories,
Eindhoven, Holland.

# BOOK REVIEW

## The Magnetic Amplifier

By J. H. REYNER. Pp. 119, with 72 illustrations and 8 plates. Stuart and Richards, 10 Bramerton Street, London, S.W.3. Price 15s.

The fairly recent development of high-quality transductors has stimulated an ever-increasing interest in transductor techniques and in useful practical applications of magnetic amplifiers. The book under review reflects the rising enthusiasm for the new subject.

The technical structure of a transductor is simple, the device being electrically and mechanically of the stature of an iron-core transformer. The theoretical treatment of its electrical performance requires, however, an unorthodox approach and non-conventional methods of thinking, even if the treatment is based on simplifying assumptions of a sweeping nature.

The object of the book under review is to explain in simple language how a transductor works and a magnetic amplifier functions, and thus to develop a background of general understanding of the broad principles of transductor design. The presentation of the subject-matter is, therefore, essentially descriptive, with numerous illustrations and few mathematical formulæ interspersed in the text.

The author has set himself to stimulate interest and to convey general ideas rather than to be rigorous. Thus any mathematical reasoning is, as far as possible, expressed in familiar terms of elementary a.c. theory.

It would be helpful for the type of reader for whom the book is intended if the presentation of the subject-matter could be slightly re-organized to eliminate one or two unnecessary repetitions of discussion and to avoid one or two irritating interruptions in the development of thought. "Magnetic field strength" is in some places meant to read "magnetic flux density," and "capacitance" or "condenser" are sometimes used instead of "capacitor."

R. F.

## BACK ISSUES

An overseas reader is anxious to obtain a copy of *Experimental Wireless & Wireless Engineer* for March 1930, Vol. 7, or a complete volume for 1930. Any reader who has either available is asked to communicate with the Publishing Dept., Iliffe & Sons, Ltd., Dorset House, Stamford St., London, S.E.1.

# ABSTRACTS and REFERENCES

Compiled by the Radio Research Board and published by arrangement with the Department of Scientific and Industrial Research

*The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.*


| | PAGE |
|---|---|
| | A |
| Acoustics and Audio Frequencies | 121 |
| Aerials and Transmission Lines | 122 |
| Circuits and Circuit Elements | 123 |
| General Physics | 127 |
| Geophysical and Extraterrestrial Phenomena | 129 |
| Location and Aids to Navigation | 130 |
| Materials and Subsidiary Techniques | 130 |
| Mathematics | 131 |
| Measurements and Test Gear | 132 |
| Other Applications of Radio and Electronics | 133 |
| Propagation of Waves | 134 |
| Reception | 135 |
| Stations and Communication Systems | 135 |
| Subsidiary Apparatus | 136 |
| Television and Phototelegraphy | 136 |
| Valves and Thermionics | 137 |
| Miscellaneous | 140 |


## ACOUSTICS AND AUDIO FREQUENCIES

534.212 **1560**

**The Propagation of Sound Waves in an Open-Ended Channel.**—W. Chester. (*Phil. Mag.*, Jan. 1950, Vol. 41, No. 312, pp. 11–33.) An exact solution is obtained for a harmonic wave propagated in the dominant mode along a channel formed by two semi-infinite parallel plates, and undergoing reflection at the open end of the channel. The reflection and transmission coefficients and the end correction are evaluated. The reflection coefficient is $\exp[-2\pi b/\lambda]$ where $2b$ is the width of the channel and $\lambda$ is the wavelength. Explicit formulae are also obtained for the change of phase in the returning wave, which is sensibly plane at large distances from the mouth, and for the transmitted energy.

534.232 : 534.321.9 : 621.3.087.6 **1561**

**Pin-Pointing Ultrasonic Energy.**—H. J. Dana & J. L. van Meter. (*Electronics*, April 1950, Vol. 23, No. 4, pp. 84–85.) Article based on 1949 National Electronics Conference paper. A magnetostriction oscillator is described which uses a thin-walled Ni-alloy tube with pointed ends in which fine holes are drilled axially. When excited at the natural frequency of about 20 kc/s, powerful air jets are expelled from the holes and produce visible changes on chemically treated paper. The ultrasonic energy concentrated in the jets is about 2 W.

534.321.9 : 534.22-14 **1562**

**Velocity of Propagation of Ultrasonic Waves in Liquid Mixtures.**—D. Sette. (*Ricerca sci.*, Nov./Dec. 1949, Vol. 19, Nos. 11/12, pp. 1338–1379.) A detailed review of work done since 1932, with a 16-page appendix tabulating results for a large number of solutions and organic liquid mixtures.

534.522.1.07 **1563**

**An Improved Schlieren Apparatus employing Multiple-Slit Gratings.**—T. A. Mortensen. (*Rev. sci. Instrum.*, Jan. 1950, Vol. 21, No. 1, pp. 3–6.) Larger fields are attained with an objective lens of given aperture by substituting gratings for the conventional knife-edge elements. In some cases a greater intensity of image illumination is also obtained.

534.6 : 621.395.623 **1564**

**The Artificial Ear of the Centre National d'Études des Télécommunications.**—P. Chavasse. (*C. R. Acad. Sci., Paris*, 12th April 1950, Vol. 230, No. 15, pp. 1390–1392.) Description of a device simulating the acoustic impedance of a normal human ear, with curves showing the concordance of measurements made with it and corresponding aural observations by 44 different individuals. See also 1058 of May.

534.6 : 621.395.623 **1565**

**Comparison between the Artificial Ears made by the** [British] **Post Office and by the Istituto Nazionale di Ultracustica.**—I. Barducci. (*Ricerca sci.*, Nov./Dec. 1949, Vol. 19, Nos. 11/12, pp. 1312–1316.) Results of tests on these ears, using both Italian and British telephone receivers, showed their characteristics to be complementary. A systematic series of such comparisons should provide a basis for standardization.

534.833.4 **1566**

**Coefficient of Acoustic Absorption for Materials made in Italy.**—M. Nuovo. (*Ricerca sci.*, Nov./Dec. 1949, Vol. 19, Nos. 11/12, pp. 1327–1331.) Graphical presentation of results for materials of various types, including those with cork, wood-pulp, vegetable- or mineral-fibre base.

621.395.61 **1567**

**The KB-3A High-Fidelity Noise-Cancelling Microphone.**—L. J. Anderson & L. M. Wigington. (*Audio Engng, N.Y.*, April 1950, Vol. 34, No. 4, pp. 16–17, 32.) A new design, providing improved discrimination against background noise.

621.395.623.7 : 621.396.621 **1568**

**High-Efficiency Loudspeakers for Personal Radio Receivers.**—Olson, Bleazey, Preston & Hackley. (*See* 1760.)

621.395.625.2 : 534.86 **1569**

**Some Problems of Disk Recording for Broadcasting Purposes.**—F. O. Viol. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 233–238.) Reprint. See 541 of March.

621.395.625.2 : 621.395.8 **1570**

**Noise Considerations in Sound-Recording Transmission Systems.**—F. L. Hopper. (*J. Soc. Mot. Pict. Televis. Engrs*, Feb. 1950, Vol. 54, No. 2, pp. 129–139.) Discussion of the unwanted noises that may be generated internally in sound-recording systems or introduced from

external sources, and of the various methods used to limit the interferences. R.f. and a.f. disturbances, cross-talk, thermal noise, shot effect, microphonics, and a.c. hum are considered.

621.395.625.3 1571

**Adjustments for Obtaining Optimum Performance in Magnetic Recording.**—A. W. Friend. (*RCA Rev.*, March 1950, Vol. 11, No. 1, pp. 38–54.) Methods of recording are discussed briefly. Using the high-frequency bias method, the harmonic distortion and noise level may be reduced considerably by the addition of d.c. bias to the recording current. The improved system is discussed at length and the results obtained under optimum operating conditions with a particular recording system are shown in numerous curves.

621.395.625.3 1572

**Magnetic Recording in Motion Pictures.**—M. Rettinger. (*Audio Engng, N.Y.*, March & April 1950, Vol. 34, Nos. 3 & 4, pp. 9–12, 35 & 28–29, 43.) "The fundamental aspects of magnetic-tape recording, particularly for motion pictures, including a description of magnetic recording, reproducing and erasing head construction, and a discussion of a.c. biasing, together with experimental results."

621.395.667 : 621.392.5.012.3 1573

**Equalizer Design Chart.**—Boegli. (*See* 1620.)

621.395.92 1574

**Miniature Electromagnetic Earpiece for Portable Hearing-Aids.**—W. Güttner. (*Z. angew. Phys.*, Feb. 1950, Vol. 2, No. 2, pp. 76–83.) Fundamental principles of design are discussed and two earpieces are described in which a Helmholtz resonator is coupled (*a*) behind (*b*) in front of the membrane, to give sensitivity in the frequency range required. See also 1335 of June.

621.396.645.029.4 1575

**For Golden Ears Only.**—Marshall. (*See* 1641.)

621.396.645.029.4 1576

**Rauland 1825 High-Fidelity Phono Amplifier.**—(*See* 1642.)

681.85 1577

**Design and Mounting of Pickup Arms.**—K. Pfeil. (*Funk u. Ton*, March 1950, Vol. 4, No. 3, pp. 140–147.) Study of common faults in alignment, and calculation of distortion produced.

681.85 1578

**A Variable-Speed Turntable and its Use in the Calibration of Disk Reproducing Pickups.**—H. E. Haynes & H. E. Roys. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 239–243.) Paper presented at the 1949 National Electronics Conference, Chicago. To overcome the variable effects caused by changes in the physical wavelength of the undulations on the conventional variable-frequency type of test record, a procedure for obtaining the frequency response of a pickup has been devised in which variations of frequency are obtained by using a constant-frequency record on a variable-speed turntable. A suitable turntable is described and results obtained with it are discussed.

534.86 1579

**Traité de Prise de Son (Treatise on Sound Pick-Up).** (Book Review).—J. Bernhart. Publishers: Editions Eyrolles, Paris, 1949, 382 pp., 2950 fr. (*Nature, Lond.*, 1st April 1950, Vol. 165, No. 4196, p. 501.) "There is ... no other text in which can be found the fruits of so varied an experience in all forms of sound pick-up, covering as it does auditorium and studio acoustics, broadcasting, the film, directive properties of typical microphones, and stereophonic reproduction."

## AERIALS AND TRANSMISSION LINES

621.315 + 621.396.67].011.2 1580

**Input Impedance of a Two-Wire Open Line and Center-Driven Cylindrical Antenna.**—T. W. Winternitz. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 299–300.) "Using potential theory, a method of analysis of the input impedance of a transmission line terminated by a cylindrical antenna is described. The results of application of this method to a particular configuration of line and antenna are presented."

621.315.212 : 621.397.5 1581

**A Multicore Television Camera Cable.**—(*Engineer, Lond.*, 27th Jan. 1950, Vol. 189, No. 4905, pp. 132–133.) A flexible cable of twenty-two conductors performing fourteen different functions. Balanced screened twin units for video and other circuits are spaced in a six-plus-one arrangement around 1.e central star-quad unit which carries the mains and 1 : 1 frame-scan currents. Six single wires are laid in the grooves between the six twins. All the units are solidly insulated. Overall diameter is 0.85 in.

621.392.26† 1582

**Waveguides** (operating) **beyond the Cut-Off Frequency: Application to Piston Attenuators.**—In the journal reference of 813 of April please read Vol. 30 for Vol. 29.

621.392.43 1583

**Improvement of the Transformation Properties of the Exponential Line by Compensation Arrangements.**—A. Ruhrmann. (*Arch. elekt. Übertragung*, Jan. 1950, Vol. 4, No. 1, pp. 23–31.) A paper prepared for publication in 1943. By connecting series capacitors and shunt inductors and resistors of suitable values at the high-impedance and low-impedance ends of an exponential line, satisfactory operation is obtained over a much wider frequency band. Three arrangements are described, giving progressively better compensation. The high-pass network equivalent of an exponential line with compensation circuits is discussed. See also 672 of 1948 (Zinke).

621.396.67 1584

**On the Effective Length of a Linear Transmitting Antenna.**—C. J. Bouwkamp. (*Philips Res. Rep.*, June 1949, Vol. 4, No. 3, pp. 179–188.) King's definition of effective length of a cylindrical centre-driven aerial with a sine distribution of current is consistent with a new general extension for any distribution of aerial current. As an example, results for a top-loaded aerial are given in tables and diagrams.

621.396.67 1585

**The Theory of *N* Coupled Parallel Antennas.**—R. King. (*J. appl. Phys.*, Feb. 1950, Vol. 21, No. 2, pp. 94–103.) The integral-equation theory of coupled aerials developed by King & Harrison (3474 of 1944), Tai (2430 of 1948) and Bouwkamp (960 of 1949) is generalized to apply to any number of units arranged symmetrically in a circle. The case of four aerials at the corners of a square is discussed in detail. Application to cage and corner-reflector aerials is indicated. The driving voltages required to maintain specified currents in *N* parallel aerials arranged in line are determined; the currents corresponding to specified voltages can only be found by solving simultaneous integral equations, but an approximate analysis is given for the special case of $\lambda/2$ dipoles.

621.396.67 1586

**Aerials protected from Effects of Local Fields.**—G. Guillner. (*Arch. elekt. Übertragung*, Feb. 1950, Vol. 4, No. 2, pp. 71–75.) Two compensation arrangements are described which greatly reduce the interference from electrical machines or other possible sources of r.f. radiation near a receiving aerial.

621.396.67 : 621.317.74.029.64 1587

**A New Type of Slotted Line Section.**—W. B. Wholey & W. N. Eldred. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 244–248.) Full paper. Summary abstracted in 2716 of 1949.

621.396.67 : 621.392.26† 1588

**The External Field Produced by a Slot in an Infinite Circular Cylinder.**—S. Silver & W. K. Saunders. (*J. appl. Phys.*, Feb. 1950, Vol. 21, No. 2, pp. 153–158.) "Expressions are derived for the external field produced by a slot of arbitrary shape in the wall of a circular wave guide (of infinite extent and infinite conductivity), the tangential components of the electric field in the slot being assumed to have been prescribed. This is accomplished by matching a Fourier representation of the external field, built up by superposition of basic sets of cylindrical waves, to a Fourier expansion of the prescribed field over the surface of the cylindrical wave guide. The far-zone field is obtained by applying the method of steepest descent to the Fourier integrals that constitute the coefficients in the expansion for the external field. The results satisfy the radiation conditions for far-zone fields."

621.396.67 : 621.392.26† 1589

**Radiation from a Transverse Slot in an Infinite Cylinder.**—C. H. Papas. (*J. Math. Phys.*, Jan. 1950, Vol. 28, No. 4, pp. 227–236.) An expression is obtained for the radiation assuming the electric field to be parallel to the slot width and its distribution along the slot to be that of the incident dominant mode. A graphical example is given of the azimuthal distribution of the field radiated from a narrow slot of length $\lambda/2$ with a cosine distribution of the electric field along it.

621.396.67 : 621.392.26† 1590

**The Radiation Characteristics of Conical Horn Antennas.**—A. P. King. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 249–251.) Measurements at wavelengths of 3 cm and 10 cm of the radiation characteristics of conical horns with waveguide excitation are in good agreement with theoretical results for absolute gain. The variations of gain and effective area with dimensions are given and also a diagram for use in plotting the radiation characteristic for conical horns of optimum design.

621.396.67 : 629.13 1591

**Shunt-Excited Flat-Plate Antennas with Applications to Aircraft Structures.**—J. V. N. Granger. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 280–287.) 1949 I.R.E. National Convention paper. Impedance/frequency characteristics of various shunt-fed plates are given, and the effect of varying the tapping point is shown. Operation is explained as an extension of folded dipole theory. Making the shunt-feed conductor part of the leading edge of the wing is suggested as aerodynamically sound. Impedance measurements on such an arrangement for 3–14 Mc/s and 90–160 Mc/s are given.

621.396.671 1592

**Gain of Aerial Systems.**—J. Brown. (*Wireless Engr*, April 1950, Vol. 27, No. 319, pp. 132–133.) Further discussion. See 287 of February and 820 of April (Bell).

## CIRCUITS AND CIRCUIT ELEMENTS

621.3.016.35 1593

**When is Nyquist's Stability Criterion Valid?**—J. Peters. (*Arch. elekt. Übertragung*, Jan. 1950, Vol. 4, No. 1, pp. 17–22.) A general criterion of stability is derived from consideration of matrix theory for active networks. Nyquist's criterion is found to agree with this general criterion only in the very important special case of no multiple negative feedback.

621.3.078 1594

**The Design of Control Circuits with the Aid of Circle Diagrams.**—W. Oppelt. (*Arch. elekt. Übertragung*, Jan. 1950, Vol. 4, No. 1, pp. 11–16.)

621.314.2 1595

**Practical Aspects of the Design of Intermediate-Frequency Transformers.**—C. E. S. Ridgers. (*J. Brit. Instn Radio Engrs*, March 1950, Vol. 10, No. 3, pp. 97–124.) The various factors involved are analysed and practical design equations are developed. Single- and double-tuned transformers and five different methods of coupling are considered. Comparisons are made between the calculated and observed performance of a particular design. Notes are included on two unconventional designs.

621.314.3† 1596

**Transductors and their Application.**—L. F. Borg. (*Elect. Times*, 23rd Feb. 1950, Vol. 117, No. 3042, pp. 269–273.) Fundamental considerations for the design of transductors are discussed and their properties examined. Suitable applications of these devices as magnetic amplifiers, rectifier controllers and voltage regulators for alternators are described and illustrated by circuit diagrams.

621.314.3† 1597

**Fundamentals of a Theory of the Magnetic Amplifier.**—W. Schilling. (*Elektrotech. Z.*, 1st Jan. 1950, Vol. 71, No. 1, pp. 7–13.)

621.314.3† 1598

**The Magnetic Amplifier.**—R. Feinberg. (*Wireless Engr*, April 1950, Vol. 27, No. 319, pp. 118–124.) "The transductor control characteristic is determined by the shape of the magnetization curve of the transductor core and by the ratio between the peak value of alternating magnetic-flux density and the value of magnetic-flux density at the knee of the magnetization curve. The control characteristic has an optimum form in regard to range of linear control, low initial load current and absence of backlash of control at zero control current if the magnetization curve has a sharp bend at the knee and high value of initial permeability, and when the peak value of alternating magnetic-flux density equals the value of magnetic-flux density at the knee of the magnetization curve."

621.314.3† 1599

**Magnetic Amplifiers.**—(*Elect. Times*, 16th March 1950, Vol. 117, No. 3045, pp. 417–419.) Summaries of and discussion on the following papers presented at an Institution of Electrical Engineers meeting:

A New Theory of the Magnetic Amplifier.—A. G. Milnes.

The Fundamental Limitations of the Second-Harmonic Type of Magnetic Modulator as applied to the Amplification of Small D.C. Signals.—F. C. Williams & S. W. Noble.

621.316.729 : 621.396.615 1600

**Theory of Synchronization by Phase Control.**—E. Labin. (*Philips Res. Rep.*, Aug. 1949, Vol. 4, No. 4, pp. 291–315. In French.) A method is described for synchronizing a self-oscillator with a pilot frequency, the oscillator frequency being governed by a control voltage applied to a reactance valve. This control voltage is derived from a mixing stage in which the phases of oscillator and pilot are compared. The method is first considered for the case where both oscillations are sinusoidal and then extended to the case where one of the two oscillations is pulsed, as in the I.G.O. system (Impulse-Governed Oscillator). Experimental results showing the practical possibilities of the method are discussed.

621.316.8 : 621.396.822

**1601**

**Dimensions and Noise of Resistors.**—A. Hettich. (*Frequenz*, Jan. 1950, Vol. 4, No. 1, pp. 14–25.) A distinction is drawn between spontaneous and current-excited noise, both types being thermal in origin, and a formula is derived for the latter in terms of the resistor length and cross-section. The constant in the expression is the "specific noise coefficient" of the material. Increasing the dimensions of a resistor of fixed value while retaining geometric similarity reduces the current-excited noise. The two classes of noise may correspond respectively with longitudinal and transverse components of electron motion.

621.318.572

**1602**

**New Application of Electronic Switches for Cathode-Ray Oscillographs.**—P. Harmegnies. (*HF, Brussels*, 1950, No. 5, pp. 117–124.) The principles of electronic switching by grid-controlled thyratrons are discussed, by which means any number of phenomena may be viewed simultaneously on a c.r.o. Such a circuit may be applied in representing different phenomena as functions of another variant, e.g., anode voltage, the coordinate axes appearing on the c.r.o. screen at the same time. An auxiliary device causes a marker trace to appear on the curve displayed. A uniform phase displacement of the synchronizing pulse moves this trace slowly along the curve so that it indicates the direction in which the curve is traced. A circuit is shown.

621.318.572

**1603**

**An Automatically Synchronized Electronic Switch.**—W. A. Budlong & B. C. Lutz. (*Rev. sci. Instrum.*, Feb. 1950, Vol. 21, No. 2, pp. 167–168.) Square waves, 180° out of phase, are synchronized with either of two input signals and are used to cut off the respective signal amplifiers alternately. Operation is satisfactory from 25 to 15 000 c/s with any input over 1.5 V.

621.392

**1604**

**A Note on the Synthesis of Electric Networks according to Prescribed Transient Response.**—M. Nadler. (*Proc. Inst. Radio Engrs*, April 1950, Vol. 38, No. 4, p. 441.) A contribution to the solution of the problem of designing coupling networks for valve amplifiers to give a required response to a step function. The effect of the grid-anode capacitance of the valves is to impose restrictions on the shape of the transient waveform if a realizable coupling network is to result. These restrictions are given in mathematical form. See also 2461 of 1949 and 3104, 3360 of 1949 (Aigrain & Williams).

621.392

**1605**

**The Synthesis of Resistor-Capacitor Networks.**—J. L. Bower & P. F. Ordung. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 263–269.) 1949 I.R.E. National Convention paper. A general method is described for deriving the balanced lattice with a prescribed output/input voltage ratio and with either unloaded output or a resistor-capacitor load. Methods of transformation to an unbalanced structure are outlined.

621.392

**1606**

**A General Review of Linear Varying Parameter and Nonlinear Circuit Analysis.**—W. R. Bennett. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 259–263.) 1949 I.R.E. National Convention paper. Variable and nonlinear systems are classified from the standpoint of their significance in communication problems. Methods of solution are reviewed and 41 relevant references are given.

621.392

**1607**

**New Time Constants for the Study of Linear Electrical Lumped-Constant Circuits.**—L. de Luca. (*Poste e Telecomunicazioni*, Dec. 1949, Vol. 17, No. 12, pp. 621–629.) Two new time constants are defined: (*a*) the time constant of integration $T$, given by $Q/I_0$, where $Q$ and $I_0$ refer to the transient current, and (*b*) the time constant of differentiation $\tau_0$, given by $-\mathfrak{I}_0/\mathfrak{I}_0'$ where $\mathfrak{I}_0$ and $\mathfrak{I}_0'$ are the initial complex current and its differential respectively. These are applied to the symbolic and characteristic equations of LCR circuits, and expressions for the current and charge at various time intervals are developed. Various examples are explained.

621.392

**1608**

**Direct Measurement of Bandwidth.**—C. R. Ammerman. (*Elect. Engng, N.Y.*, March 1950, Vol. 69, No. 3, pp. 207–212.) The equivalent bandwidth of a network is defined as $f_2 - f_1 = \int_0^\infty (A_f^2/A_a^2)\,df$, where $A_f$ is the amplification of the network, a function of frequency, and $A_a$ is the maximum value of $A_f$. The bandwidth may be found directly by comparing the network under test with a standard network. A noise voltage is applied to the two networks and the two output voltages are measured, from which the required bandwidth is obtained in terms of the ratio of the maximum output voltages of the two networks at their respective resonance frequencies, with input voltage constant, and the known bandwidth of the standard. The equipment is described and practical details of the method are given, with typical experimental results.

621.392.001.11

**1609**

**Remarks on the Basis of Network Theory.**—R. M. Redheffer. (*J. Math. Phys.*, Jan. 1950, Vol. 28, No. 4, pp. 237–258.) Discussion of certain general theorems relative to transmission and reflection in transmission lines and networks.

621.392.001.11 : 621.3.015.3

**1610**

**Applications of Network Theorems in Transient Analysis.**—Y. P. Yu. (*J. Franklin Inst.*, Nov. 1949, Vol. 248, No. 5, pp. 381–398.) An account of the application of selected theorems (Thévenin's, equivalent current-generator, compensation and superposition) to the simplification of the solution of problems involving transients in electrical networks. In each case examples illustrating the method are given.

621.392.26† : 621.396.611.39

**1611**

**A Note on Coaxial Bethe Hole Directional Couplers.**—E. L. Ginzton & P. S. Goodwin. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 305–309.) Perfect directivity is theoretically obtainable at any wavelength if the diameter of the coupling hole is small compared with $\lambda/8$, but with such small holes the coupling is too weak for many applications. Increasing the coupling by enlarging the hole affects both s.w.r. and directivity, but these can be restored to their proper values over a large frequency range by neutralizing the series inductance caused by the hole by means of added lumped capacitance.

621.392.43

**1612**

**A Broadband Transition from Coaxial Line to Helix.**—C. O. Lund. (*RCA Rev.*, March 1950, Vol. 11, No. 1, pp. 133–142.) The transition section is a cylindrical conductor with a helical slot of pitch varying from infinity to that of the helical line, with a coaxial cylindrical outer conductor. An approximate theory for the section is outlined, based on an impedance concept for helical lines and the theory of tapered transmission lines. Transitions of reasonable length can be constructed with a voltage s.w.r.<1.5 over a very wide band.

621.392.43

**1613**

**The Wide-Band Problem in Decimetre-Wave Technique.**—H. H. Meinke. (*Elektrotechnik, Berlin*, May 1948, Vol. 2, No. 5, pp. 137–142.) Discussion of

methods of compensating reactive elements to obtain minimum frequency dependence in the case of coupling circuits and wide-band aerials. Different circuit arrangements and various types of wide-band transformer are described.

1614

621.392.43

**Impedance-Matching Networks.**—R. O. Rowlands. (*Wireless Engr*, April 1950, Vol. 27, No. 319, pp. 113–118.) "A method is described of designing a four-terminal network whose image impedance at one pair of terminals is a constant resistance and at the other pair of terminals a complex quantity varying with frequency."

1615

621.392.5

**Dissipative Quadripoles with Dual Iterative Impedance.**—R. Possenti. (*Poste e Telecomunicazioni*, April 1949, Vol. 17, No. 4, pp. 225–230.) In certain cases a quadripole may have two physically realizable iterative impedances, only one of which is a pure reactance.

1616

621.392.5

**Quadripoles.**—M. Leroy. (*Bull. Soc. franç. Elect.*, March 1950, Vol. 10, No. 102, pp. 128–134.) Discussion of the problem of the synthesis of 4-terminal networks. An alternative method to that of Gewertz (1934 Abstracts, p. 514) is suggested.

1617

621.392.5

**Frequency Analysis of Variable Networks.**—L. A. Zadeh. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 291–299.) 1949 I.R.E. West Coast Convention paper. An approach to the analysis of linear networks with time-dependent impedance parameters is made by a frequency-domain technique in which a transfer factor or system function, which is a function of time, is used, as in the analysis of nonvariable networks. Two methods of deriving the system function are considered, the first based on Schelkunoff's wave-perturbation method and the second, a much simpler method, for use with slowly-varying parameters only. The two methods are applied to a low-pass bandwidth-modulated RC half-section, the same result being obtained in both cases for conditions in which the second method is applicable.

1618

621.392.5

**Investigation of the Effect of Resistive Damping on the Phase Shift in Lattice Networks.**—W. Taeger. (*Funk u. Ton*, March & April 1950, Vol. 4, Nos. 3 & 4, pp. 107–118 & 193–198.) Damping factor ($\delta$) and phase constant are calculated for networks with small and with large losses. A phase change only occurs in the pass-band when ohmic resistance may be neglected, and in the suppression band when circuit elements are lossy. Small resistive damping, up to $\delta = 0.1$, has only slight effect on the phase shift in a simple lattice network, but where L and C appear together in one of the branches slight damping alters the characteristics considerably.

1619

621.392.5

**Attenuation and Delay Equalizers.**—W. R. Lundry. (*Elect. Engng, N.Y.*, March 1950, Vol. 69, No. 3, p. 216.) Summary of A.I.E.E. Fall Meeting paper, 1949, discussing equalizers for coaxial lines such as those of the New York/Chicago television circuit. Since such long lines may require about 30 equalizers, manufacturing tolerances must necessarily be small and provision must be made for individual adjustments of inductance and transmission loss.

1620

621.392.5.012.3 : 621.395.667

**Equalizer Design Chart.**—C. P. Boegli. (*Electronics*, April 1950, Vol. 23, No. 4, p. 114.) Bass and treble attenuation or accentuation of two types of RC equalizers are easily determined and curves thus obtained resemble those obtained by more exact computation.

1621

621.392.5.018.1 : 621.396.615

**The Design of RC Oscillator Phase-Shifting Networks.**—T. S. Dutton : W. R. Hinton. (*Electronic Engng*, April 1950, Vol. 22, No. 266, pp. 162–164.) Comment on 840 of April and author's reply.

1622

621.392.52

**Admittance and Transfer Function for an $n$-Mesh RC Filter Network.**—E. W. Tschudi. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 309–310.) The coefficients in the admittance and transfer functions of an $n$-mesh low-pass filter are exactly calculated as functions of $n$. If these functions are true for a particular value of $n$, they are also true for the value $(n + 1)$ and also when $n = 1$.

1623

621.392.52.092

**Phase-Shift Calculation.**—W. Saraga & J. Freeman. (*Wireless Engr*, Jan. 1950, Vol. 27, No. 316, pp. 32–33.) Discussion of various simple methods of deriving an approximation to the curve showing the image phase-shift coefficient of a multi-section filter as a function of frequency.

1624

621.392.52.092

**Phase-Shift Calculation.**—N. O. Johannesson. (*Wireless Engr*, April 1950, Vol. 27, No. 319, p. 133.) A more exact method of calculation than that described by Saraga & Freeman (1623 above), for the particular case where all minima in the image attenuation are equal to one another. Simple formulae give easier calculation with little loss of accuracy.

1625

621.396.611.3

**Resonant Frequencies and Characteristics of a Resonant Coupled Circuit.**—C. L. Cuccia. (*RCA Rev.*, March 1950, Vol. 11, No. 1, pp. 121–132.) The characteristics are derived and charted so that they may be used in predicting the performance of u.h.f. valves using an external resonant circuit for tuning, f.m., and control.

1626

621.396.611.4 : 537.533 : 621.396.619

**Interaction of a Spiral Electron Beam and a Resonant Microwave Cavity.**—F. K. Willenbrock & S. P. Cooke. (*J. appl. Phys.*, Feb. 1950, Vol. 21, No. 2, pp. 114–125.) The particular cavity considered here is of the cylindrical TE11 type in a steady axial magnetic field. If the cavity is excited in a linearly polarized mode, the electromagnetic field will drive the electrons in a helical trajectory with an expanding radius, and the electrons will excite and transfer energy to a degenerate mode oriented spatially at right angles to the driving field. In the driving plane of polarization (both planes of polarization if the cavity is excited in a circularly polarized mode), the electron beam will excite a field in phase opposition to the driving field in a manner analogous to the counter e.m.f. in an electromechanical generator. The converse case of a TE11 cavity excited by a spiral beam of electrons is also considered. Possible applications in microwave generators and modulators are mentioned.

1627

621.396.615

**High-Stability Oscillator.**—G. G. Gouriet. (*Wireless Engr*, April 1950, Vol. 27, No. 319, pp. 105–112.) The factors governing frequency stability are analysed and a general expression for the stability of a series-resonant circuit is derived. The practical arrangement discussed is essentially a Colpitts oscillator in which a series LC circuit is substituted for the tuning inductance. Performance figures are given for the circuit used for many years as a variable-frequency driver for B.B.C. s.w. transmitters. See also 2193 of 1948 (Clapp).

621.396.615 **1628**

**Phase-Shift Oscillator.**—W. C. Vaughan. (*Wireless Engr*, April 1950, Vol. 27, No. 319, p. 129.) Corrections to paper abstracted in 324 of February.

621.396.615 : 621.396.611.21 **1629**

**A High-Stability Quartz Oscillator.**—In the journal reference of 2107 of May please read Vol. 30 for Vol. 29.

621.396.615.11 **1630**

**Frequency Stability and Parasitic Oscillations in a Bridge-Stabilized Low-Frequency Oscillator.**—H. Matthes. (*Frequenz*, Jan. & Feb. 1950, Vol. 4, Nos. 1 & 2, pp. 1–10 & 41–50.) Circle diagrams of the loop amplification are used to study methods of suppressing unwanted oscillations in Meacham's oscillator (3717 of 1938 & 263 of 1939), which incorporates an amplitude-limiting bridge circuit and combined feedback. Variations of the transformer tuning, imperfect phasing of feedback circuits and peculiarities of valve characteristics can cause spurious oscillations. Formulae are derived for calculating these effects. Suppression of oscillations above the nominal frequency (1 kc/s in the case considered) may lead to lower-frequency disturbance if the by-pass capacitances are too small. Bridge-stabilized oscillators with only one transformer may be somewhat more convenient; the design of wide-band circuits for this case is considered.

621.396.615.17 **1631**

**The Miller Circuit as a Low-Speed Precision Integrator.**—I. A. D. Lewis. (*Electronic Engng*, April 1950, Vol. 22, No. 266, p. 141.) Corrections to paper abstracted in 1111 of May.

621.396.619.23 **1632**

**Balanced Rectifier Modulators without Transformers.**—D. G. Tucker. (*Electronic Engng*, April 1950, Vol. 22, No. 266, pp. 139–141.) Two types of modulator are described, a shunt and a series type. An approximate mathematical analysis of their performance is given and a note on the carrier (or switching) circuit is included.

621.396.619.23 **1633**

**Non-Linear Effects in Rectifier Modulators.**—V. Belevitch. (*Wireless Engr*, April 1950, Vol. 27, No. 319, pp. 130–131.) A more detailed analysis, taking account of the finite conductances of the rectifiers and the finite carrier-source resistance. The results of the analysis are outlined for (*a*) a ring modulator, (*b*) a Cowan modulator with an auxiliary rectifier connected in opposite polarity, (*c*) a single Cowan modulator. See also 2184 of 1949 and 1552 of June (Tucker & Jeynes).

621.396.621.54 : 621.396.622.6 **1634**

**On Mixing with Detectors.**—A. Klemt. (*Frequenz*, Feb. 1950, Vol. 4, No. 2, pp. 50–54.) Discussion of general design principles of h.f. mixer circuits. In crystal detectors, if the back resistance is large enough, the same relations hold as for diodes having the same slope. This back resistance should not fall below about 30 kΩ for mixing with the oscillator fundamental, or below about 100 kΩ for mixing with the 2nd or 3rd harmonics. Maximum amplification in the mixer stage occurs for an oscillator voltage of 1–2 V. Circuit arrangements for measurement of amplification, resistance, and equivalent noise resistance are shown.

621.396.645 **1635**

**On the Theory of Semi-Distributed Amplifiers.**—J. L. Steinberg. (*Onde élect.*, March 1950, Vol. 30, No. 276, pp. 121–127.) For the same bandwidth the classical feedback amplifier requires fewer valves, but the distributed amplifier has a much lower noise factor. For a given bandwidth the noise factor of a distributed unit with $n$ identical valves decreases as $n$ is increased, but approaches a limit when $n = 6$. The semi-distributed amplifier described consists of a low-pass distributed input stage followed by a cascade m.f. amplifier. With such an amplifier it is possible to construct u.h.f. radiation meters whose performance is comparable with that of optical pyrometers.

621.396.645 **1636**

**Transmission Factor of Differential Amplifiers.**—D. H. Parnum. (*Wireless Engr*, April 1950, Vol. 27, No. 319, pp. 125–129.) A cathode-coupled input stage is used to illustrate the theory. The effect of an anti-phase signal applied between the two grids of the stage cancels in the common cathode resistor and output at full gain is obtained between the anodes. An interfering in-phase signal applied between earth and the two grids in common produces negative feedback, which cuts the in-phase output; but if the halves of the stage are not identical there will be an anti-phase output from this input. The ratio, anti-phase gain/in-phase gain, is thus insufficient to define performance; a knowledge of the "transmission factor", defined as the ratio of (*a*) anti-phase gain due to anti-phase input, to (*b*) anti-phase gain due to in-phase input, is also required. A method of controlling the value of this factor is explained and recommendations are made for the design of a successful amplifier. See also 680 of 1948 (Johnston).

621.396.645 : 539.16.08 **1637**

**Noise in Ionization-Chamber Pulse Amplifiers.**—R. Wilson. (*Phil. Mag.*, Jan. 1950, Vol. 41, No. 312, pp. 66–76.) The general problem of the detection of heavy ionization particles in the presence of $\gamma$ radiation and amplifier background noise is considered. The dependence of signal-noise ratio on the shape of the initial current pulse is discussed. For optimum signal/noise ratio the pulse should have a certain exponential form. A circuit suitable for the production of such pulses is described.

621.396.645 : 681.142 **1638**

**Analog Computers for Servo Problems.**—McDonald. (*See* 1715.)

621.396.645.012.3 **1639**

**Graphical Methods for the Study of Cathode-Follower Circuits.**—G. Salmet. (*Onde élect.*, March 1950, Vol. 30, No. 276, pp. 128–139.) Cathode-voltage/cathode-current characteristics analogous to those for anode loading afford a means of rapid determination of optimum conditions for cathode-follower operation. Secondary phenomena such as those caused by an unbalanced input can also be treated by this method, which applies equally for complex loads, as in the case of a video-frequency amplifier with a very large pass-band.

621.396.645.029.3 **1640**

**The Amplification of Very-Low Frequency Electric Currents.**—G. Lehmann. (*Onde élect.*, April 1950, Vol. 30, No. 277, pp. 169–177.) Review of different methods.

621.396.645.029.4 **1641**

**For Golden Ears Only.**—J. Marshall. (*Audio Engng, N.Y.*, April 1950, Vol. 34, No. 4, pp. 13–15, 32.) Description of a high-fidelity audio amplifier, including circuit diagram and component values. All components are standard commercial types. The circuit consists of a single-ended input stage transformer-coupled to a push-pull triode driver stage, again transformer-coupled to a push-pull triode output stage. Feedback is applied to each stage. The amplifier is capable of delivering 10 W over the range 20–20 000 c/s with less than 1% distortion. Below 8 W the distortion is too low to be measured accurately. Intermodulation distortion is approximately 1% at 10 W and 2% at 15 W.

621.396.645.029.4 **1642**

**Raulard 1825 High-Fidelity Phono Amplifier.**—(*Audio Engng, N.Y.*, April 1950, Vol. 34, No. 4, p. 25.) Includes a preamplifier which may be attached to the main amplifier or mounted separately.

621.396.645.029.4 **1643**

**Selective RC Audio-Frequency Amplifier.**—K. Feher & G. Kurtze. (*Frequenz*, March 1950, Vol. 4, No. 3, pp. 72–76.) Selectivity is achieved by use of a feedback circuit comprising a Wien-Robinson bridge, tuning being accomplished by adjustment of R and C in the bridge arms. In an amplifier illustrated, the frequency range is 25 c/s to 30 kc/s and amplification about $5 \times 10^5$. The bridge circuit may be switched out for wide-band operation with the same amplification.

621.396.645.029.4 **1644**

**Audio-Frequency Selective Amplifiers.**—S. W. Punnett. (*J. Brit. Instn Radio Engrs*, Feb. 1950, Vol. 10, No. 2, pp. 39–59.) Three types of frequency-discriminator feedback circuit are considered, the simple RC series-parallel arrangement, the twin-T network and the Wien bridge. The analysis covers the effective $Q$, bandwidth, gain variations and stability as functions of frequency. When constancy of the $Q$-factor of an amplifier is of primary importance, the twin-T network should be used, but it has the disadvantage of requiring variation of three components, which must be accurately ganged. When some variation of $Q$ is permissible, the simple circuit with only two variable elements is preferable, being both cheaper and easier to align.

621.396.645.029.4 : 52 : 621.396.822 **1645**

**Selective Amplification at Low Frequencies.**—In the journal reference of 1120 of May please read Vol. 30 for Vol. 29.

621.396.645.37 **1646**

**On the Power Gain and the Bandwidth of Feedback Amplifier Stages.**—A. van der Ziel & K. S. Knol. (*Philips Res. Rep.*, June 1949, Vol. 4, No. 3, pp. 168–178.) The effects of feedback on the power gain $g$ and bandwidth $B$ of an amplifier stage are treated theoretically. When the stability limit is reached for a given feedback value, $g$ remains finite and for correct output coupling can only have very high values when $B$ is very small. Application of the formulae to a grounded-grid triode stage indicates that small output losses can give a large decrease of $g$. Values of $Bg$ are plotted against $g$ for various values of feedback capacitance and output losses.

621.396.813 **1647**

**The Nonlinearity of Valve Circuits due to the Voltage-Dependence of Valve Input Capacitance.**—F. Feil. (*Arch. elekt. Übertragung*, Feb. 1950, Vol. 4, No. 2, pp. 65–70.) In consequence of the space-charge effect, the input capacitance of a valve, and hence the impedance of the valve circuit, varies with the alternations of the applied voltage. In a screen-grid valve with a slope of 10 mA/V the capacitance variation may be 0.5 to 1 pF per 1 V change at the grid. Calculated values of the nonlinear distortion agree with those measured.

534.01 + 535.56 **1648**

**Theory of Oscillations.** [Book Review]—A. A. Andronow & C. E. Chaikin. Publishers: Princeton University Press, Princeton, N.J., 1949, 336 pp., $6.00. (*Proc. Inst. Radio Engrs*, April 1950, Vol. 38, No. 4, p. 449.) Published originally in Russia in 1937 and now appearing as an edited translation by a group at Princeton University. A presentation of "the basic mathematics relating to the theory of nonlinear oscillatory systems". The book is recommended to all those who are interested in learning more about the analysis of such systems and the differential equations that describe them.

621.3.015.3 **1649**

**Transients in Linear Electrical Systems.** [Book Review]—K. A. Krug. Publisher: Gosenergoizdat, Moscow, 1948, 344 pp., 19 roubles. (*Bull. Acad. Sci. U.R.S.S. tech. Sci.*, Nov. 1949, No. 11, pp. 1741–1744. In Russian.)

621.392 **1650**

**Communication Circuits.** [Book Review]—L. A. Ware & H. R. Reed. Publishers: J. Wiley & Sons, New York, 1949, 396 pp., $5.00. (*Proc. Inst. Radio Engrs*, April 1950, Vol. 38, No. 4, p. 449.) "A very good book intended as first-course material for all students of communication engineering, regardless of the frequency range with which they will be concerned. The material is limited to lumped networks and transmission lines. No active circuits are included. For the field covered, the choice of material is good."

621.392.52 **1651**

**Einführung in die Siebschaltungstheorie der elektrischen Nachrichtentechnik (Introduction to the Theory of Filter Circuits for Electrical Communications).** [Book Review]—R. Feldtkeller. Publishers: S. Hirzel Verlag, Stuttgart, 160 pp., 12 DM. (*Wireless Engr*, April 1950, Vol. 27, No. 319, p. 134.) This volume does not deal with high frequencies, which are treated in the author's Theorie der Rundfunksiebschaltungen (584 of 1946). "The treatment throughout is very thorough and the book is undoubtedly a valuable addition to the literature of the subject."

621.396 **1652**

**Electron-Tube Circuits.** [Book Review]—S. Seely. Publishers: McGraw-Hill Book Co., New York, 320 pp., $6.00. (*Radio & Televis. News, Radio-Electronic Engng Supplement*, April 1950, Vol. 14, No. 4, p. 28.) "This college text is the outgrowth of several courses organized by the author on electron-tube circuits and applications that covered many of the important circuits in use during the second world war." Circuits for performing mathematical operations and those developed in connection with radar applications are discussed in greater detail than in most textbooks of to-day.

## GENERAL PHYSICS

53.081 + 537.712 **1653**

**Proposals and Recommendations concerning the Definitions and Units of Electromagnetic Quantities.**—P. Cornelius. (*Philips Res. Rep.*, June 1949, Vol. 4, No. 3, pp. 232–237.) A set of definitions of e.m. quantities and the corresponding units which follow the adoption of the rationalized Giorgi system, with suggestions that may be useful in the transfer from one of the older c.g.s. systems. See also 867 of April (Cornelius & Hamaker).

534.26 + 535.42 **1654**

**On Certain Integral Equations in Diffraction Theory.**—J. W. Miles. (*J. Math. Phys.*, Jan. 1950, Vol. 28, No. 4, pp. 223–226.) For diffraction of a plane wave by an infinite slit, the integral equation defining the field in the slit is given, with solution, and also its Fourier transform and solution. Corresponding equations are obtained for diffraction by a ribbon. These results lead to useful integral relations between the Mathieu functions.

535.8 **1655**

**On the Aberrations of the Field-Flattened Schmidt Camera.**—E. H. Linfoot & P. A. Wayman. (*Mon. Not. R. astr. Soc.*, 1949, Vol. 109, No. 5, pp. 535–556.) A general aberration theory is developed and applied to a comparison of the optical performances of the field-flattened and the plain Schmidt cameras in a typical special case.

537.226.2 : 546.431.82 **1656**

**Dielectric Constants and Polarizabilities of Ions in Simple Crystals and Barium Titanate.**—S. Roberts. (*Phys. Rev.*, 15th Oct. 1949, Vol. 76, No. 8, pp. 1215–1220.)

537.52 **1657**

**Plasma and the Langmuir Layer : On the Theory of Electrical Probes in Gas Discharges.**—F. Wenzl. (*Z. angew. Phys.*, Feb. 1950, Vol. 2, No. 2, pp. 59–75.)

537.533 : 538.569 : 621.396.822 **1658**

**Fluctuation Phenomena arising in the Quantum Interaction of Electrons with High-Frequency Fields.**—D. K. C. MacDonald & R. Kompfner. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, p. 304.) Correction to 603 of March.

538.114 **1659**

**Theory of Magnetic After-Effect for Massive Materials in the Rayleigh Domain.**—L. Néel. (*J. Phys. Radium*, Feb. 1950, Vol. 11, No. 2, pp. 49–61.) Extension to massive materials of theory previously given for fine-grain materials. See also 3151 and 3152 of 1947, 2782 and 3159 of 1949.

538.114 **1660**

**Laws Relating to Magnetic After-Effect.**—L. Libboutry. (*C. R. Acad. Sci., Paris*, 13th March 1950, Vol. 230, No. 11, pp. 1042–1044.) In weak fields, the after-effect resulting from a variation of the magnetic field may be measured by applying a small additional variation of the same sign; this variation should have a certain minimum value. The phenomena are discussed with reference to Néel's theory (1659 above).

538.114 **1661**

**Magnetic After-Effect in the Rayleigh Domain.**—J. C. Barbier. (*C. R. Acad. Sci., Paris*, 13th March 1950, Vol. 230, No. 11, pp. 1040–1041.) Experimental study of massive materials shows that the decrease of the square root of the remanent magnetization is proportional to the time elapsed from the suppression of the field, and is independent of the value of this field and the duration of its application. These results are in very good agreement with Néel's theory (1659 above).

538.24 **1662**

**Practical Calculation of Magnetizing Force.**—A. A. Halacsy. (*Proc. Instn elect. Engrs*, Part I, March 1950, Vol. 97, No. 104, pp. 37–42.) Any vector component of the magnetizing force at a point in space due to an electric current is directly related to a vector representing the surface swept over by the radius vector of length $1/\sqrt{r}$ with its origin and direction the same as those of the radius vector $r$ directed from the point in question to consecutive points on the path of the influencing current. This relation enables magnetizing force, distribution of the magnetic field, electromagnetic forces, etc., to be calculated by means of simple geometry in all practical cases, including those in which no simple analysis has so far been possible.

538.56 **1663**

**Fresnel's Laws for Centimetre Waves.**—J. C. Simon. (*C. R. Acad. Sci., Paris*, 17th April 1950, Vol. 230, No. 15, pp. 1386–1388.) A note forming the introduction to a more comprehensive paper on diffraction effects obtained with perforated metal plates. Analogies with classical optical properties are confirmed experimentally, in particular, Brewster's law is verified.

538.56 **1664**

**On the Focusing of a Wave.**—J. C. Simon. (*C. R. Acad. Sci., Paris*, 24th April 1950, Vol. 230, No. 17, pp. 1513–1514.) Further discussion of the analogy between radio and light waves (1132 of May). In the case of nonisotropic media the analogy is rather with crystal optics.

538.56 **1665**

**Graphical Study of the Dispersion of Electro-Magneto-Ionic Waves.**—V. A. Bailey & J. A. Roberts. (*Aust. J. sci. Res., Ser. A.*, Sept. 1949, Vol. 2, No. 3, pp. 307–321.) A graphical method is given which leads to an approximate solution of the dispersion equation. The following information is readily obtained:— (*a*) the frequency bands in which undamped waves or wave-groups can grow as they progress, (*b*) the wave-number bands in which unattenuated waves can grow in time, (*c*) the phase and group velocities, refractive indices, and attenuation and damping coefficients, (*d*) the effect of collisions between electrons and other particles on the attenuation of a wave or wave-group.

538.56 : 537.71 **1666**

**Concerning 'The Impedance of Free Space'.**—In the journal reference of 876 of April please read Vol. 30 for Vol. 29.

538.561 : 537.533 : [537.29+538.6 **1667**

**Generation and Amplification of Waves in Dense Charged Beams under Crossed Fields.**—O. Buneman. (*Nature, Lond.*, 25th March 1950, Vol. 165, No. 4195, pp. 474–476.) Using plane electrodes without resonators or corrugations, a dense beam of charged particles may be made to produce oscillations similar to those in magnetrons, by a process which is purely electrostatic and can occur at low fields and low frequencies. The conclusions of a theoretical study of this phenomenon are discussed and expressions are given for the electric field distribution, the transverse particle-velocity distribution, the wave-velocity along the beam, and, in the case of a feedback system, the resonance frequency and amplification. The system may have value as a simple generator or amplifier for microwaves and the mechanism may possibly explain the production of solar noise. See also 3412 of 1949 (Haeff).

538.569.4+538.561.029.65 **1668**

**Microwave Spectroscopy in the Region from Two to Three Millimeters.**—O. R. Gilliam, C. M. Johnson & W. Gordy. (*Phys. Rev.*, 15th April 1950, Vol. 78, No. 2, pp. 140–144.) A description of the technique employed for spectroscopic measurements on gases such as CO at millimetre wavelengths. Wavelengths down to 1.96 mm have been generated by frequency multiplication using a silicon crystal and a klystron source.

538.569.4 **1669**

**Measurements on the Absorption of Microwaves : Part 4 — Non-Polar Liquids.**—D. H. Whiffen. (*Trans. Faraday Soc.*, Feb. 1950, Vol. 46, No. 326, pp. 124–130.) Experimental study and discussion of the small dielectric losses which occur in $C_6H_6$ and $CCl_4$ and in liquids with similar symmetrical molecules.

538.569.4 **1670**

**Measurements on the Absorption of Microwaves: Part 5 — The Variation of Relaxation Time with Solvent.**—D. H. Whiffen. (*Trans. Faraday Soc.*, Feb. 1950, Vol. 46, No. 326, pp. 130–136.) Description of a method of measurement of dielectric losses at 1.6-cm wavelength. Experimental results are given for 25 systems of polar molecules dissolved in non-polar solvents.

538.66 **1671**

**Action of a Periodic Magnetic Field on a Thin Spherical Metal Film.**—A. Colombani. (*C. R. Acad. Sci., Paris*, 20th March 1950, Vol. 230, No. 12, pp. 1149–1150.) Calculation of heating effects, based on electromagnetic potentials.

549.211 : 537.533.9 : 537.311.3 **1672**

**Electron Bombardment Conductivity in Diamond: Part 2.**—K. G. McKay. (*Phys. Rev.*, 15th March 1950, Vol. 77, No. 6, pp. 816–825.) Previously published results have been revised by use of an improved alternating-field method of internal space-charge neutralization. Lower limits are set for the mobilities of electrons and positive holes at room temperature. Measurements of the decay of current due to internal space-charge fields are in reasonable agreement with theory. A double-pulse technique is used to study the rates of release of positive holes and electrons from traps.

537 + 538 **1673**

**Principles of Electricity and Electromagnetism.** [Book Review]—G. P. Harnwell. Publishers: McGraw-Hill Book Company, New York, 2nd edn, 670 pp., $6.00. (*Rev. sci. Instrum.*, Jan. 1950, Vol. 21, No. 1, pp. 90–91.) A comprehensive textbook and valuable reference book for practising physicists and for electrical engineers whose work borders on physics. Much of the text has been rewritten. New topics added include electron structure of crystals, magnetic induction accelerators, nuclear induction, v.h.f. oscillators, the propagation of e.m. waves in metallic enclosures.

537.226 **1674**

**Theory of Dielectrics.** [Book Review]—H. Fröhlich. Publishers: Clarendon Press, Oxford, 1949, 179 pp., $4.50 or 18s. (*Rev. sci. Instrum.*, Jan. 1950, Vol. 21, No. 1, p. 90: *Trans. Faraday Soc.*, Nov. 1949, Vol. 45, No. 323, p. 1084.) "A systematic account of the theory of dielectric constant and of dielectric loss. The treatment, based on classical statistical mechanics, is rigorous and methodical . . . The book can be recommended without reserve, not only to specialists in the field of dielectrics but also to all physicists, chemists and applied scientists who wish to gain a clear and comprehensive grasp of this important field."

## GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

521.15 : 538.12 **1675**

**The Equations of Codazzi and the Relations between Electromagnetism and Gravitation.**—A. Gião. (*Phys. Rev.*, 15th Sept. 1949, Vol. 76, No. 6, pp. 764–768.) See also 3128 of 1949.

523.53 : 621.396.9 **1676**

**A Radio Echo Method for the Measurement of the Heights of the Reflecting Points of Meteor Trails.**—Clegg & Davidson. (*See* 1692.)

523.7 : 538.122 **1677**

**The Sun's General Magnetic Field.**—P. A. Sweet. (*Mon. Not. R. astr. Soc.*, 1949, Vol. 109, No. 5, pp. 507–516.) The effects of large-scale convection currents and differential angular velocity on the solar magnetic field are studied in detail.

523.72 : 621.396.822 **1678**

**Transients in an Ionized Medium with Applications to Bursts of Solar Noise.**—J. C. Jaeger & K. C. Westfold. (*Aust. J. sci. Res., Ser. A*, Sept. 1949, Vol. 2, No. 3, pp. 322–334.) Mathematical solutions of a number of transient problems in linear propagation in a homogeneous ionized medium are given. Discussion of the effects of collisions, inhomogeneity of medium, and magnetic field is included. If a localized disturbance occurs in the solar atmosphere where the collision frequency is $\nu$ and the frequency of plasma oscillations is $\omega_0$, radiation will be emitted on all frequencies greater than $\omega_0$ with intensity determined by the nature of the disturbance, and with damping determined by $\nu$. There is a small difference between the arrival times on different frequencies, and for each frequency there is a direct wave and an echo wave a few seconds later. Many of the phenomena of bursts of solar noise on frequencies between 85 Mc/s and 19 Mc/s are consistent with these predictions.

551.1 + 550.38 **1679**

**The Earth's Interior and Geomagnetism.**—W. M. Elsasser. (*Rev. mod. Phys.*, Jan. 1950, Vol. 22, No. 1, pp. 1–35.) A review of present knowledge of the earth's interior, excluding the extensive information available concerning the crust, with particular reference to seismic data, chemical composition, mechanical and thermal properties and magnetic phenomena.

551.510.5 **1680**

**Cellular Atmospheric Waves in the Ionosphere and Troposphere.**—D. F. Martyn. (*Proc. roy. Soc. A*, 22nd March 1950, Vol. 201, No. 1065, pp. 216–234.) Horizontally travelling disturbances of electron densities in the $F_2$ region, and the supposed vertically travelling disturbances of Wells, Watts & George (3279 of 1946) are both explainable as horizontally travelling atmospheric cellular waves, the theory of which is developed. These waves also appear to be the cause of the microbarometric oscillations in the troposphere. In the ionosphere, the earth's magnetic field affects the ionic motion and makes the horizontally travelling disturbances simulate vertically moving electron clouds. Consideration of the boundaries for the cellular waves indicates a value of $\gamma$, the ratio of the specific heats of air, considerably less than 1.4 in the $F_2$ region.

551.510.535 **1681**

**On the Diurnal Variations of Electron Concentration of the $F_2$ Layer at Equatorial Stations.**—R. Eyfrig. (*Naturwissenschaften*, Feb. 1950, Vol. 37, No. 3, pp. 67–68.) The mean hourly values of $f^0(F_2)$ for two equatorial stations with about 180° difference in longitude, Huancayo and Bandoeng, for March and September 1944 show a difference which is not explained by Appleton's temperature effect. The curve for Palau, March 1944, is similar to that for Huancayo. The magnetic inclinations of these two stations are + 0.3° and − 0.5°, that of Bandoeng being about − 32°. The flattening of the curve around noon is attributed to the influence of the geomagnetic field.

551.510.535 : 523.745 **1682**

**The Effect of the Sun on the Normal E Layer of the Ionosphere.**—E. Harnischmacher. (*C. R. Acad. Sci., Paris*, 27th March 1950, Vol. 230, No. 13, pp. 1301–1302.) Previous investigations have indicated that in Chapman's law for the variation of critical frequency with the sun's zenithal angle, the index $n$ should be higher than $\frac{1}{4}$. Analysis of more recent measurements leads to a formula giving values of $n$ lying between 0.29 and 0.32 in latitude 48°, and between 0.33 and 0.36 at the equator if the mean monthly relative sunspot number varies between 0 and 150.

551.510.535 : 551.542 **1683**

**A Correlation between Ionospheric Phenomena and Surface Pressure.**—M. W. Jones & J. G. Jones. (*Phys. Rev.*, 15th March 1950, Vol. 77, No. 6, p. 845.) Observations at College, Alaska, indicate that oscillations in the F layer are approximately in opposite phase to the oscillations of the air near the surface of the earth, as measured by the variations of barometric pressure.

551.510.535 : 621.396.11 **1684**

**Ionospheric Recordings.**—A. Bolle, S. Silleni & C. A. Tiberio. (*Ann. Geofis.*, July 1949, Vol. 2, No. 3, pp. 377–387.) Systematic ionosphere soundings were resumed at the Istituto Nazionale di Geofisica in August 1948. Graphs of $F_2$ critical frequencies for each month up to March 1949 are given, and items of special interest discussed briefly. These include a few examples

of triple magneto-ionic splitting. As reported by Newstead (2505 of 1948) they were accompanied by strong sporadic-E ionization and preceded by a fall in critical frequency.

551.577 : 621.396.9 **1685**

**An Investigation of the Dimensions of Precipitation Echoes by Radar.**—J. R. Mather. (*Bull. Amer. met. Soc.*, Oct. 1949, Vol. 30, No. 8, pp. 271–277.)

551.577 : 621.396.9 **1686**

**The Rain Required for a Radar Echo.**—D. W. Perrie. (*Bull. Amer. met. Soc.*, Oct. 1949, Vol. 30, No. 8, pp. 278–281.)

551.594 **1687**

**Observations of the Electric Field of the Atmosphere at Monaco.**—J. Rouch. (*C. R. Acad. Sci., Paris*, 24th April 1950, Vol. 230, No. 17, pp. 1485–1487.) The mean value of 269 measurements made during 1949 at altitude 70 m is 366 V/m. The field exhibits a seasonal variation, with a maximum in winter and minimum in summer. Meteorological conditions are given for the 3 occasions when the field was >1 000 V/m.

551.594.5 **1688**

**Radar Reflections from Auroras.**—P. A. Forsyth, W. Petrie, F. Vawter & B. W. Currie. (*Nature, Lond.*, 8th April 1950, Vol. 165, No. 4197, pp. 501–502.) Systematic search for radar echoes on frequencies of 3 kMc/s and 106.5 Mc/s has been carried out at Saskatoon, Canada. Echoes have been observed regularly on the lower frequency, but not at all on the higher one, during auroral displays. Whether the reflections are from the lower and brighter parts of the auroras or from levels a few kilometres above them cannot at present be decided, as the 106.5-Mc/s equipment could only be rotated about a vertical axis.

## LOCATION AND AIDS TO NAVIGATION

621.396.9 **1689**

**The Port Radar at Antwerp.**—(*H.F., Brussels*, 1950, No. 5, p. VII.) Brief notice of experimental radar equipment for the river Scheldt. The scanning aerial is about 15 m above river level and the range is adjustable between 40 and 2 miles. Several scanning towers may be installed at different points along the river banks with a central observation station or, alternatively, only one or two very high towers with special equipment may be used.

621.396.9 **1690**

**Radar for the Merchant Navy.**—M. Gilli. (*Poste e Telecomunicazioni*, April 1949, Vol. 17, No. 4, pp. 233–236.) An account of radar technique and equipment from the standpoint of current British and American practice.

621.396.9 **1691**

**Volume Scanning with Conical Beams.**—D. Levine. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 287–290.) 1949 I.R.E. National Convention paper. "The equations of motion for a radar scan that provides a nearly constant number of pulses to all point targets are derived. In addition, the periods for this scan and for the more conventional one having constant angular velocities are presented. For these types of scanner it is found that the optimum cross-over point between adjacent beams is at the 2.1-db point of the antenna pattern."

621.396.9 : 523.53 **1692**

**A Radio Echo Method for the Measurement of the Heights of the Reflecting Points of Meteor Trails.**—J. A. Clegg & F. A. Davidson. (*Phil. Mag.*, Jan. 1950, Vol. 41, No. 312, pp. 77–85.) The range is determined from 60-Mc/s pulse echoes obtained by broadside reflection from the meteor trail and the elevation of the reflecting point is found by comparing the amplitudes of the echo signals picked up by two aerials mounted at different heights (usually $\lambda/2$ and $3\lambda/4$) above a plane, horizontal, perfectly reflecting surface. Results obtained during the Quadrantid shower of 1949 are in good agreement with visual observations.

621.396.9 : 629.13.052] : 621.396.6.001.2 **1693**

**Mechanical Development of a Radio Altimeter.**—Croft. (*See* 1837.)

621.396.933 **1694**

**Pictorial Display in Aircraft Navigation and Landing.**—L. F. Jones, H. J. Schrader & J. N. Marshall. (*Proc. Inst. Radio Engrs*, April 1950, Vol. 38, No. 4, pp. 391–400.) According to the 15-year plan prepared by the U.S.A. Radio Technical Commission for Aeronautics, a pictorial situation display is to be provided at the ground station and in aircraft, and is to be used in the traffic-control zone. Displays for traffic control and instrument landing are illustrated. In traffic control the picture can be used in various ways, which are discussed, with particular reference to the teleran system. Developments are described concerning self-identification in the picture, the use of graphechon storage tubes, altitude coding, and picture brightness. Both further technical development and psychological investigation are necessary before optimum display methods can be determined.

## MATERIALS AND SUBSIDIARY TECHNIQUES

537.226.2 : 546.431.82 **1695**

**Dielectric Constants and Polarizabilities of Ions in Simple Crystals and Barium Titanate.**—S. Roberts. (*Phys. Rev.*, 15th Oct. 1949, Vol. 76, No. 8, pp. 1215–1220.)

537.311.31 : 538.221 **1696**

**Improvements and Recent Results relating to the Measurements of the Temperature Variation of the Resistance of Ferromagnetic Materials.**—G. Mannevy-Tassy. (*C. R. Acad. Sci., Paris*, 20th March 1950, Vol. 230, No. 12, pp. 1150–1152.)

538.221 **1697**

**Study of Static and Dynamic Magnetostriction Phenomena of Austenitic Fe/Ni Alloys.**—H. Devèze. (*Ann. Phys., Paris*, Jan.–Feb. 1950, Vol. 5, pp. 80–144.) A thesis in four parts. The different apparatus used for static and dynamic measurements is described, particular attention being given to temperature stability. For measurement of elongation under a direct magnetizing force the optical interference method was preferred, as being both accurate and rapid. For dynamic measurements the radiation from the end of the ferromagnetic rod energized by a variable-frequency 400-W ultrasonic generator was directed upon a quartz microphome. Static magnetostriction is studied as a function of nickel content in a series of cylindrical rods and also in 4 invar alloys subjected to different heat treatments. The theory of dynamic magnetostriction is developed to determine the amplitude of oscillations. Experimental results are analysed and optimum operating conditions are established.

538.221 **1698**

**The Preparation and Magnetic Properties of Ferrites of Manganese and Cobalt.**—C. Guillaud & H. Creveaux. (*C. R. Acad. Sci., Paris*, 27th March 1950, Vol. 230, No. 13, pp. 1256–1258.)

538.221 **1699**

**Ferromagnetic Properties of Mixed Ferrites of Cobalt and Zinc and of Manganese and Zinc.**—C. Guillaud & H. Creveaux. (*C. R. Acad. Sci., Paris*, 17th April 1950, Vol. 230, No. 16, pp. 1458–1460.)

538.244.2 : 538.12

1700

**The Magnetic Field inside a Solenoid.**—J. R. Barker. (*Brit. J. appl. Phys.*, March 1950, Vol. 1, No. 3, pp. 65–67.) Convenient formulae are derived for computing the extent of the region of uniform field inside both thick and thin coils, and an estimate of their accuracy is made. A set of tables is included to simplify the calculations.

539.23 : 537.311.31

1701

**Dependence on Applied Potential of the Resistance of Very Thin Deposited Metal Films at Low Temperatures.**—N. Mostovetch & B. Vodar. (*C. R. Acad. Sci., Paris*, 6th March 1950, Vol. 230, No. 10, pp. 934–936.) Resistivity measurements for films of various metals in the temperature range 293°–2°K show marked departures from Ohm's law at low temperatures. A direct effect of the field on the conduction electrons is indicated.

546.431.82 : 621.315.612.4

1702

**The Dielectric Properties of Barium Titanate at Low Temperatures.**—R. F. Blunt & W. F. Love. (*Phys. Rev.*, 15th Oct. 1949, Vol. 76, No. 8, pp. 1202–1204.)

546.431.82 : 621.315.612.4

1703

**Domain Structures and Phase Transitions in Barium Titanate.**—P. W. Forsbergh, Jr. (*Phys. Rev.*, 15th Oct. 1949, Vol. 76, No. 8, pp. 1187–1201.)

546.431.82 : 621.315.612.4

1704

**The Electric and Optical Behavior of $BaTiO_3$ Single-Domain Crystals.**—W. J. Merz. (*Phys. Rev.*, 15th Oct. 1950, Vol. 76, No. 8, pp. 1221–1225.)

548.0 : 537.228.1

1705

**Synthetic Crystals at Ultrasonic Frequencies.**—C. E. Green. (*Electronics*, April 1950, Vol. 23, No. 4, pp. 110–112.) Short discussion of the relative advantages of ADP and Rochelle-salt crystals as compared with quartz, for the frequency range 10–150 kc/s. Ease of production may for some applications offset their large temperature coefficients of frequency.

549.514.51

1706

**Quartz, its Treatment and its Use in Telecommunications Technique.**—R. Sueur, J. Norbert, P. Andrieux & M. Cornebise. (*Onde Élect.*, Feb. & March 1950, Vol. 30, Nos. 275 & 276, pp. 67–75 & 140–148.) Illustrated description of production technique, methods of mounting and applications in oscillators and filters.

549.514.51 : 536.42

1707

**The $\alpha\beta$ Transformation Point of Quartz indicated by the Dielectric Constant.**—J. P. Pérez. (*C. R. Acad. Sci., Paris*, 6th March 1950, Vol. 230, No. 10, pp. 932–933.) Capacitance/temperature curves for quartz-plate capacitors exhibit kinks or discontinuities at the transformation point near 570°C.

621.315.61

1708

**Fluid Dielectrics for the Decimetre and Centimetre Wave Band.**—W. Endres & H. Köhler. (*Frequenz*, March 1950, Vol. 4, No. 3, pp. 57–63.) Discussion of composite dielectrics, which are of importance as offering a continuous range of values of dielectric constant. Measurements of the dielectric constant for different combinations of powdered ceramic materials with air or transformer oil confirm Lichtenecker & Rother's logarithmic law of mixtures.

621.315.612.4

1709

**Properties of Calcium-Barium Titanate Dielectrics.**—E. N. Bunting, G. R. Shelton & A. S. Creamer. (*Bur. Stand. J. Res.*, Sept. 1949, Vol. 43, No. 3, pp. 237–244.) Properties measured were shrinkage, water absorption, linear expansion, dielectric constant and power factor. The electrical properties were measured at 0.05, 1, 20 and 3 000 Mc/s, and the highest values of dielectric constant were found to be associated with high baria content. The properties of some specimens changed with time. The expansion coefficient of all specimens was relatively high and of the order of $10 \times 10^{-6}$ per 1°C for the temperature range 25°–700°C.

778.3 : 621.396.9

1710

**Single-Pulse Recording of Radar Displays.**—L. C. Mansur. (*Tele-Tech*, Jan. 1950, Vol. 9, No. 1, pp. 30–33, 54.) Techniques are described which were originally developed for investigating fluctuations of fixed targets and their effect on the operation of moving-target indicators. Typical records are reproduced and various applications of the method are suggested.

535.37

1711

**Fluorescence and Phosphorescence.** [Book Review]—P. Pringsheim. Publishers: Interscience Publishers, New York and London, 1949, 794 pp., 120s. (*Nature, Lond.*, 29th April 1950, Vol. 165, No. 4200, pp. 659–660.) "Provides a clear, balanced, fair and complete account of present knowledge . . . an indispensable reference book for the student who requires either an overall picture of the subject or detailed information on any particular part."

## MATHEMATICS

517.912.2

1712

**A Note on the Numerical Integration of Differential Equations.**—W. E. Milne. (*Bur. Stand. J. Res.*, Dec. 1949, Vol. 43, No. 6, pp. 537–542.) An integration method for ordinary differential equations is developed, in which the approximation formulae contain derivatives of higher order than those contained in the differential equation itself. The method is particularly useful for linear differential equations. Numerical examples are given for Bessel's differential equation.

519.283 : 621.39.001.11

1713

**A Simplified Derivation of Linear Least-Square Smoothing and Prediction Theory.**—H. W. Bode & C. E. Shannon. (*Proc. Inst. Radio Engrs*, April 1950, Vol. 38, No. 4, pp. 417–425.) The principal results of the Wiener-Kolmogoroff smoothing and prediction theory for stationary time series are developed by a new method. The mathematical procedure has, for the most part, a direct physical interpretation based on electric circuit theory and does not involve integral equations or the autocorrelation function. The cases treated are the "infinite lag" smoothing problem, pure prediction in the absence of noise, and the general smoothing prediction problem with noise present. The basic assumptions of the theory are discussed in order to clarify the question as to when the theory will be appropriate and to avoid possible misapplication.

681.142 : 621-526

1714

**Electronics and Experimental Mathematics.**—In the journal reference of 919 of April please read Vol. 30 for Vol. 29.

681.142 : 621.396.645

1715

**Analog Computers for Servo Problems.**—D. McDonald. (*Rev. sci. Instrum.*, Feb. 1950, Vol. 21, No. 2, pp. 154–157.) Exact equations are derived for the operational amplifiers used in analogue computers. By making certain approximations, simpler equations are obtained which indicate that in the case of servomechanisms the operational amplifiers may be of simpler type, with lower gain, than for more general problems.

501

1716

**Applied Mathematics for Engineers and Scientists.** [Book Review]—S. A. Schelkunoff. Publishers: D. van Nostrand, 1948, 472 pp., 36s. (*Phil. Mag.*, Jan. 1950, Vol. 41, No. 312, p. 108.) "The first half treats the general mathematical methods and the second half introduces the frequently occurring transcendental

functions. The portion dealing with differential equations is particularly well written. The introductions to the use of Green's function and the Laplace transform method are two of the most valuable features of the book. The principles involved have been illustrated by concrete examples as far as possible. A chapter on tensors and matrices would perhaps have increased the general usefulness of the book."

## MEASUREMENTS AND TEST GEAR

621.317.3 : 621.385.2 : 621.315.59 — 1717

**Applications of Germanium Crystal Diodes in Precision-Measurement Technique.**—A. Perlstain. (*Bull. schweiz. elektrotech. Ver.*, 28th May 1949, Vol. 40, No. 11, pp. 337–354. In German.) Basic circuits and the development of bridge and star modulator systems and their variants are considered. Detailed descriptions are given of (*a*) arrangements for the measurement of frequencies, voltage mean values, and capacitance, (*b*) a method for point-to-point curve plotting, (*c*) equipment for vector measurements, (*d*) a method for harmonic analysis of voltage curves. Auxiliary apparatus includes a circuit for producing square waves without using valves. Accuracy of measurement is indicated in each case and temperature effects are discussed. The indicator generally used is a d.c. galvanometer in a compensation circuit. The methods described can be used up to 200 kc/s.

621.317.329 : 537.291 — 1718

**An Automatic Plotter for Electron Trajectories.**—D. B. Langmuir. (*RCA Rev.*, March 1950, Vol. 11, No. 1, pp. 143–154.) A scale model of the e.s. field is produced in an electrolyte trough. The voltages picked up by a pair of wire probes determine the radius of curvature of the trajectory simulating that of an electron. Arrangements are provided so that when started from a selected point in a given initial direction the radius of curvature of the path of the probes is automatically adjusted so that they trace a simulated trajectory. Mechanical design, electrical circuits, and performance are discussed.

621.317.33 — 1719

**Methods for Obtaining the Voltage Standing-Wave Ratio on Transmission Lines Independently of the Detector Characteristics.**—A. M. Winzemer. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 275–279.) By means of the basic transmission line equations and the general law of detection, expressions are derived which give the voltage s.w.r. as a function of measurable electrical angles, the expressions being independent of the response law of the detector. Special cases of the general equations are discussed covering applications where high or low s.w.r. values are to be determined, with or without the use of the voltage maxima or minima. The extension of the methods to the case of attenuating transmission lines is considered.

621.317.336.029.64 — 1720

**Directional Coupler for Impedance Measurement at U.H.F.**—R. Masson-Genon & P. Brissonneau. (*C. R. Acad. Sci., Paris*, 27th March 1950, Vol. 230, No. 13, pp. 1258–1259.) The impedance to be measured terminates a waveguide to which is coupled, by means of a directional coupler of attenuation about 20 db, a section of waveguide terminated at one end by a sliding piston and at the other by an accurately matched crystal detector, the current in which varies with the position of the piston. The unknown impedance is readily found from measurements of the maximum and minimum currents. Accuracy is of the same order as when using a movable probe, but the construction of the apparatus is much simpler.

621.317.373.029.64 — 1721

**Experimental Determination of Phase Characteristics of Centimetric Waveband Circuits.**—M. Denis & P. Palluel. (*Microtecnic, Lausanne*, Sept. 1949–Feb. 1950, Vol. 3, Nos. 5 & 6, & Vol. 4, No. 1, pp. 218–224, 289–294 & 3–9.) English version of 682 of March.

621.317.74.029.64 : 621.396.67 — 1722

**A New Type of Slotted Line Section.**—W. B. Wholey & W. N. Eldred. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 244–248.) Full paper. Summary abstracted in 2716 of 1949.

621.317.755 : 621.396.645 — 1723

**A Wide-Band Amplifier for Oscilloscopes.**—(*Radio & Electronics, Wellington, N.Z.*, 1st March 1950, Vol. 5, No. 1, pp. 18–19.) A simple 5-valve circuit giving a balanced output from an unbalanced input and having a response useful up to 4.5 Mc/s. It is suitable for examination of square or irregular waveforms and will fill the screen of a 5-inch c.r. tube from an input of 1 V r.m.s.

621.317.772 — 1724

**Wide-Band Audio Phasemeter.**—J. R. Ragazzini & L. A. Zadeh. (*Rev. sci. Instrum.*, Feb. 1950, Vol. 21, No. 2, pp. 145–149.) A stabilized feedback amplifier is arranged to produce a 90° phase shift and its input is added to a variable fraction of its output to give a resultant e.m.f. of variable phase. This calibrated phase shift is matched to the unknown phase difference, using a c.r.o. as indicator. The error of measurement is $< 0.5°$ for frequencies from 20 c/s to 100 kc/s.

621.317.772 — 1725

**A New Measuring Instrument for Time-Delay Measurements.**—W. Dillenburger. (*Frequenz*, Jan. 1950, Vol. 4, No. 1, pp. 10–13.) Circuit details and description of a phase meter for the frequency range 100 kc/s to 20 Mc/s. Calibration is simple. Indication is direct and single-valued between 0° and 360° and within reasonable limits is independent of the intermediate frequency used in the meter and of the amplitude of the input signals.

621.317.784 — 1726

**Dynamometer Type Wattmeter for Audio Frequencies.**—D. Alvey. (*Electronic Engng*, April 1950, Vol. 22, No. 266, pp. 146–148.) The basic principles of operation at a.f. are summarized and expressions for frequency errors when used as a voltmeter, ammeter or wattmeter are derived. A range of instruments covers measurements of voltage to 300 V, current to 3 A and power to 500 W. Various applications are noted.

621.317.794 — 1727

**Measurement of Small Powers at Centimetre Wavelengths.**—J. Broc. (*Onde Elect.*, March 1950, Vol. 30, No. 276, pp. 108–120.) Detailed account of methods and apparatus noted in 2860 of 1949.

621.317.794 : 621.317.733 — 1728

**Analysis of the Direct-Current Bolometer Bridge.**—D. M. Kerns. (*Bur. Stand. J. Res.*, Dec. 1949, Vol. 43, No. 6, pp. 581–589.) "The behavior of a direct-current Wheatstone bridge with a nonlinear element (bolometer) in one arm, as used in the measurement of microwave power, is analyzed. Results of rather general applicability are obtained by ascribing arbitrary values to the resistances making up the bridge-network and by employing a resistance law of a general form for the bolometer element. Emphasis is placed upon the development of first-order theory in convenient form. General characteristics of the behavior of the type of network (and nonlinear element) under consideration are indicated. Results obtained include expressions for the derivatives of galvanometer current with respect to radio-frequency power, ambient temperature, source-voltage, and source-resistance."

621.319.4 1729

**The Temperature Coefficient of Standard Air Capacitors.**—G. Zickner. (*Elektrotechnik, Berlin*, May 1948, Vol. 2, No. 5, pp. 147–152.) The factors which determine the coefficient are analysed and methods of reducing it are discussed. Descriptions are given of different designs of fixed and variable capacitors in which the temperature coefficient is much reduced by use of materials with different expansion coefficients in a "grid-iron" type of construction. Physical constants of different conducting and insulating materials are tabulated.

621.397.6.001.4 + 621.397.6 1730

**The Video-Frequency Stage.**—Gondry. (*See* 1782.)

## OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.232 : 534.321.9 : 621.3.087.6 1731

**Pin-Pointing Ultrasonic Energy.**—Dana & van Meter. (*See* 1561.)

539.10.08 1732

**Small Probing Geiger-Müller Counters.**—C. V. Robinson. (*Rev. sci. Instrum.*, Jan. 1950, Vol. 21, No. 1, pp. 82–84.) Experiments with mixtures of argon and ethyl acetate in small counters are described. Construction details are given for two brain-probing counters of diameter 3 mm and 2 mm respectively.

621.3.083.7 : 621.396 1733

**High-Accuracy Radio-Telemetry.**—I. Podliasky & J. Zakheim. (*C. R. Acad. Sci., Paris*, 27th March 1950, Vol. 230, No. 13, pp. 1260–1262.) In a system for measuring the velocity of a moving object such as an aircraft, in which differently phased beats are produced at a series of ground stations receiving signals from both the moving object and a ground transmitter, it is important to keep the difference between the two signal frequencies constant. Servomechanisms are provided at the ground transmitter for this purpose. Instrumental accuracy to within about 1 part in $10^8$ is possible, the overall accuracy being limited only by variation of propagation velocity in the vicinity of the ground.

621.317.39 1734

**Electrical Device for the Calculation of Product Integrals.**—O. Schäfer & G. Lander. (*Arch. elekt. Übertragung*, Feb. 1950, Vol. 4, No. 2, pp. 59–64.) Curves representing the two functions concerned are scanned simultaneously by oscillating light beams. The resulting outputs from the two photocells used are applied, after amplification, to an integrator. The time for a complete scan is about 2 minutes.

621.365.54† 1735

**H.F. Heating speeds Steel Forging.**—(*Elect. Times*, 9th March 1950, Vol. 117, No. 3044, pp. 355–362.) Description of press-forge plant at Bromsgrove, Worcestershire. For another account see 420 of February.

621.38 : 629.13 1736

**Electronics in Aircraft Design.**—A. L. Whitwell. (*J. Brit. Instn Radio Engrs*, March 1950, Vol. 10, No. 3, pp. 79–95.) A survey of electronic methods used in measurements on aircraft structures, both on the ground and in flight; strain, displacement, velocity, acceleration and pressure are dealt with, and multi-channel apparatus developed for recording high-frequency vibrations is described.

621.384.6 1737

**The Cyclosynchrotron.**—M. L. Oliphant. (*Nature, Lond.*, 25th March 1950, Vol. 165, No. 4195, pp. 466–468.) A first model is under construction for installation in the School of Physical Sciences in the Australian National University, Canberra. Protons are accelerated in a synchrocyclotron and a synchrotron guiding field is produced about the final orbit in such a way that the passage from one type of acceleration to the other is smooth. Final particle energies of the order of 2–3 kMeV should be obtained.

621.384.611.2† 1738

**Design Study for a 3 000-MeV Proton Accelerator.**—M. S. Livingston, J. P. Blewett, G. K. Green & L. J. Haworth. (*Rev. sci. Instrum.*, Jan. 1950, Vol. 21, No. 1, pp. 7–22.) Discussion of a proton synchrotron, to be called a cosmotron, now under construction at Brookhaven National Laboratory, Upton, N.Y.

621.384.611.2† 1739

**Betatron-Starting in Electron Synchrotrons.**—J. J. Wilkins. (*Phil. Mag.*, Jan. 1950, Vol. 41, No. 312, pp. 34–53.)

621.385.15.001.8 1740

**Recent Applications of Electron Multiplier Tubes.**—Allen. (*See* 1815.)

621.385.833 1741

**The Practice of Electron Microscopy.**—(*J. R. micr. Soc.*, March 1950, Vol. 70, Part I, No. 349, pp. 1–141.) A comprehensive treatise compiled by the Electron Microscopy Group of the Institute of Physics and edited by D. G. Drummond. General optical considerations and basic techniques are discussed and methods are described for the examination of small particles, thin-film specimens, surface replicas, micro-organisms, and biological tissues. Special techniques, the calibration of magnification and resolving power, and photographic technique are also considered. Illustrative photographs and an extensive bibliography are included.

621.385.833 1742

**A Three-Stage Electron Microscope with Stereographic Dark Field, and Electron Diffraction Capabilities.**—M. E. Haine, R. S. Page & R. G. Garbitt. (*J. appl. Phys.*, Feb. 1950, Vol. 21, No. 2, pp. 173–182.) A continuous range from ×1 000 to ×100 000 is covered by use of an intermediate projector lens.

621.385.833 1743

**Principal Rays for a Family of Electron Lenses and Mirrors.**—É. Regenstreif. (*C. R. Acad. Sci., Paris*, 27th March 1950, Vol. 230, No. 13, pp. 1262–1264.)

621.398 1744

**Mains Control Installations: Report on the S.E.V. (Schweizerischer Elektrotechnischer Verein) Conference in Berne, Dec. 1949.**—(*Bull. schweiz. elektrotech. Ver.*, 4th March 1950, Vol. 41, No. 5, pp. 153–200.) Introductory papers by E. Erb (in German), reviewing possible methods of control, and by M. Roesgen (in French), outlining the requirements of centralized remote control systems, are followed by: Research and New Productions of the Compagnie des Compteurs, by J. Pelpel (in French): description of different relays for carrier current control systems and consideration of optimum frequency; Principal Features of the Landis & Gyr Centralized Control System, by W. Koenig (in German): discussion of factors determining the choice of frequency for the pulse-width method; Mains Control Installations of the Zellweger A.G., Uster, by O. Grob (in German): description of a system using stored energy for the control signals; The Sauter Remote Control System, by E. Spahn (in German): pulse system operating on 2 kc/s and using a triode rectifier and synchronous selector at the receiving end. The concluding discussion (in German) includes descriptions of other systems and equipment.

629.13.055.6 : 621.317.733.011.4 1745

**Airplane Fuel Gage.**—C. R. Schafer. (*Electronics*, April 1950, Vol. 23, No. 4, pp. 77–79.) The liquid fuel in the aircraft tank forms the dielectric in a tubular capacitor in one arm of a bridge, the unbalance voltage operating a servo-motor which restores the bridge balance and indicates fuel level mechanically. Advantages of the equipment are ruggedness, freedom from slosh effects, good accuracy, and indication of weight rather than volume of fuel.

621.317.083.7 : 621.396 1746

**Principles and Methods of Telemetering.** [Book Review.]—P. A. Borden & G. M. Thynell. Publishers: Chapman & Hall, London, 1948, 230 pp., 27s. (*Beama J.*, Oct. 1949, Vol. 56, No. 145, p. 549.) "A work which presents a technically analytical and descriptive treatment of a large variety of telemetering devices. . . . It is an informative record of U.S. practice in telemetering."

621.38 1747

**Elements of Electronics.** [Book Review.]—G. Windred. Publishers: Chapman & Hall, London, 1949, 185 pp., 15s. (*Beama J.*, Feb. 1950, Vol. 57, No. 152, pp. 46–47.) "An elementary, non-mathematical introduction to the subject, giving a survey of principles and applications.

## PROPAGATION OF WAVES

538.56 1748

**Vectorial Space of Regular Waves outside a Closed Surface.**—J. Brodin. (*C. R. Acad. Sci., Paris*, 12th April 1950, Vol. 230, No. 15, pp. 1388–1390.) The general expression of Huyghens' principle for a regular monochromatic wave outside a closed surface *S* and at infinity is given in the form of a series of waves produced by the basic layers of electric and magnetic dipoles tangential to *S*. The development involves arbitrary coefficients permitting its application to particular problems. See also 523 and 712 of March.

538.56 1749

**Particular Case of Huyghens' Principle.**—J. Brodin. (*C. R. Acad. Sci., Paris*, 3rd April 1950, Vol. 230, No. 14, pp. 1345–1347.) Extension of 712 of March to the case in which the solutions of the integral equation are not zero. See also 1218 of May.

538.566.2 1750

**The Geometric Locus of the Coefficient of Reflection of Electromagnetic Waves due to a Discontinuity in the Gradient of the Dielectric Constant, in the case of Small Gradient.**—G. Eckart. (*C. R. Acad. Sci., Paris*, 13th March 1950, Vol. 230, No. 11, pp. 1044–1045.)

621.396.11 1751

**Velocity of Light and of Radio Waves.**—L. Essen. (*Nature, Lond.*, 15th April 1950, Vol. 165, No. 4198, pp. 582–583.) Recent measurements of the velocity of light and radio waves are discussed. The value of 299 776 ± 4 km/sec quoted by Birge (1941) for light has been generally accepted, but the measurements described suggest that this figure is too low. Careful measurements with the Shoran radio system in America give a value of 299 792 ± 2.4 km/sec after making a refractive-index correction. Essen & Gordon-Smith have obtained a value of 299 792 km/sec, with an estimated maximum error of ± 9 km/sec, by measurements in vacuo of the resonance frequency of a metallic cavity of known dimensions. A later figure given by Essen is 299 792.5 ± 3 km/sec. Recent optical determinations by Bergstrand in Sweden using a Kerr cell and photocell give a value of 299 792.7 ± 0.25 km/sec.

621.396.11 1752

**Some Remarks on the Ionospheric Double Refraction: Part 2.**—H. Bremmer. (*Philips Res. Rep.*, June 1949, Vol. 4, No. 3, pp. 189–205.) Maxwell's equations for plane wave propagation through a stratified anisotropic medium are reduced to a set of four simultaneous first-order equations. A general method is indicated for deriving W.K.B. approximations from these equations and details are given for the simplified case of an isotropic medium. The intensities of the waves reflected by the stratified anisotropic medium are discussed and results combined with the saddle-point method of evaluating exponential integrals to find the geometric-optical approximation for the field strength after one ionospheric reflection. Part 1: 3519 of 1949.

621.396.11 1753

**Ionosphere Data Deduced from Direct Tests on Radiotelegraphic Links in the Italian Army Network.**—S. Silleni. (*Ann. Geofis.*, July 1949, Vol. 2, No. 3, pp. 388–399.) Results for the first three months of a year's survey of communications on 7.695 Mc/s. Values of the critical frequencies and their distribution were deduced from the data collected. Comparison with results obtained from ionosphere soundings showed fairly good agreement for mean values; distribution about the mean was more scattered in the case of the links by a factor which remained almost constant over the period considered.

621.396.11 : 551.510.535 1754

**Ionospheric Predictions and Radiocommunication.**—M. Nicolet. (*HF, Brussels*, 1950, No. 5, pp. 125–142.) General discussion of the factors influencing the propagation of short waves in the ionosphere. Methods for predicting usable frequencies and ionospheric conditions are explained. The theories underlying prediction calculations are outlined.

621.396.11 : 551.510.535 1755

**Ionospheric Recordings.**—Bolle, Silleni & Tiberio. (*See* 1684.)

621.396.11.029.6 1756

**Reflection of Radio Waves from a Rough Sea.**—L. V. Blake. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 301–304.) "At microwave frequencies, the sea cannot always be assumed to be a smooth, or mirror-like, reflecting surface. It is shown that when the sea is rough the reflected field will be a randomly fluctuating one, or will at least have a fluctuating component, even though the radiated signal is of constant amplitude and frequency. The central limit theorem of mathematical probability theory is used to derive the probability-density functions for the amplitude of the reflected signal, and for the amplitude of the combined reflected and direct-path signals. The possible practical significance of these results is discussed."

621.396.11.029.64 1757

**A Theory of Radio Scattering in the Troposphere.**—H. G. Booker & W. E. Gordon. (*Proc. Inst. Radio Engrs*, April 1950, Vol. 38, No. 4, pp. 401–412.) "The theory of scattering by a turbulent medium is applied to scattering of radio waves in the troposphere. In the region below the horizon of the transmitter, energy is received (*a*) by diffraction round the curved surface of the earth (modified as appropriate by atmospheric refraction), and (*b*) by scattering from turbulence in the region of high field strength above the horizon. At distances beyond the horizon that are not too great, we may think of (*a*) as giving the mean signal received, and (*b*) as giving the fading. However, contribution (*b*) usually decreases with distance more slowly than

contribution (a). Beyond a certain distance, therefore, contribution (b) becomes predominant and the mean signal is no longer given by (a)." A graphical comparison of field strength according to the scattering theory (using expected values of atmospheric turbulence and departure of the refractive index from its mean value) with experimental results obtained over sea at a wavelength of 9 cm shows that the theory is fully adequate to explain the high field strengths observed at great distances.

621.396.81 1758

**Approximate Formula for the Night Field-Intensity Curves established by the C.C.I.R.**—C. Glinz. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st April 1950, Vol. 28, No. 4, pp. 147–151. In French.) The Committee's work in producing these curves is reviewed. Weyrich's theory (1928 Abstracts, p. 516) on radiation from a finite aerial between two perfectly conducting planes is applied to the propagation of waves of wavelength 200–2 000 m between earth and the ionosphere. Neglecting distance attenuation, the result of the effects of interference and superposition is a cylindrical wave with field varying as $1/\sqrt{r}$, where $r$ is the transmission distance. The inclusion of damping according to Sommerfeld's function $f(p)$ leads to formulae which for large values of $r$ serve as an approximation for the field intensity curves.

## RECEPTION

621.396.621 1759

**Frequency Conversion by Phase Variation.**—G. Diemer & K. S. Knol. (*Philips Res. Rep.*, June 1949, Vol. 4, No. 3, pp. 161–167.) The highest theoretical conversion transconductance $g_{cm}$ is obtained when the valve transconductance has its optimum positive value and the phase angle decreases linearly by $2\pi$ radians during one cycle of the oscillator voltage. With these conditions $g_{cm}$ equals the maximum h.f. transconductance and improves by a factor $\pi/2$ on the value of $g_{cm}$ given by Herold's phase-reversal method. Experimentally the electron beam is modulated using for each voltage two systems giving deflections at right angles to each other and in phase quadrature, the i.f. current being taken in balance from two concentric annuli.

621.396.621 : 621.395.623.7 1760

**High-Efficiency Loudspeakers for Personal Radio Receivers.**—H. F. Olson, J. C. Bleazey, J. Preston & R. A. Hackley. (*RCA Rev.*, March 1950, Vol. 11, No. 1, pp. 80–98.) The direct radiator, combination direct radiator and acoustical phase inverter, horn, and combination horn and phase inverter types of loudspeaker are discussed. The measured response frequency characteristic of each type is compared with the theoretical response. An efficiency of 25% has been obtained with a combination of horn and phase inverter; this system has been incorporated in a complete 4-valve receiver of total volume only 25 cubic inches.

621.396.621.5 1761

**Dual-Diversity Frequency-Shift Reception.**—D. G. Lindsay. (*Proc. Instn Radio Engrs, Aust.*, Feb. 1950, Vol. 11, No. 2, pp. 29–42.) A survey of frequency-shift systems and their advantages is followed by a general description of a receiver for 1.5–30-Mc/s teleprinter service, using the audio-filter method of separating mark and space signals (2 125 and 2 975 c/s respectively). Practical filter and limiter characteristics and typical performance figures for the receiver, limiter-detector, relay, frequency-deviation indicator and voice-frequency keying unit are given.

621.396.622 : 621.397.62 1762

**Continuous Tuning for TV Sets.**—B. K. V. French. (*FM-TV*, March 1950, Vol. 10, No. 3, pp. 9–11, 33.) Illustrated details and performance data of an input tuning system covering the range 54–216 Mc/s and permitting reception of f.m. broadcasts on 88–108 Mc/s without the addition of any other components. Another advantage of continuous as compared with step-by-step tuning is that necessary adjustments to compensate for aging do not require a serviceman. Variable-inductance tuning is used because of the wide range to be covered; a new inductuner developed for this purpose is described.

621.396.828 1763

**Radio Interference Suppression.**—S. F. Philpott. (*Electrician*, 31st March 1950, Vol. 144, No. 3746, pp. 1025–1029.) Interference caused by small motors is the main form considered, and means of reducing the interference voltages to $< 0.5$ mV at the equipment terminals are discussed; various arrangements of capacitors and chokes being shown. The interference voltages remaining after applying different methods of suppression to a small universal motor are tabulated. A set designed by the Radio Branch of the British Post Office for the measurement of asymmetric r.f. voltages between the terminals of a machine and earth is briefly described.

## STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 1764

**Recent Developments in Communication Theory.**—C. E. Shannon. (*Electronics*, April 1950, Vol. 23, No. 4, pp. 80–83.) A non-mathematical treatment of communication systems, discussing methods of coding information, its measurement, and the capacity of a channel in relation to bandwidth, noise and power. Ideal and practical systems are briefly considered; a speech-transmission method being suggested which requires a bandwidth of only 2.3 c/s on a channel with a signal/noise ratio of 20 db.

621.39.001.11 : 519.283 1765

**A Simplified Derivation of Linear Least Square Smoothing and Prediction Theory.**—Bode & Shannon. (*See* 1713.)

621.395.5 : 621.395.661.2 1766

**Note on Matching the Impedance of Terminal Equipment to that of Modern Long-Distance Telephone Lines.**—R. Sueur. (*Ann. Télécommun.*, March 1948, Vol. 3, No. 3, pp. 105–108.) Matching defects produce additional crosstalk on the symmetrical lines used for multichannel carrier-current telephony. Equipment for long-distance coaxial cables should be matched to the lines for the lowest frequencies transmitted. C.C.I.F. recommendations for long-distance lines are mentioned and considered with reference to the transmission of television signals.

621.396.41 : 621.39.001.11 1767

**Theoretical Aspects of Asynchronous Multiplexing.**—W. D. White. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 270–275.) 1949 I.R.E. National Convention paper. Recent work in communication theory by Shannon (1361 and 1649 of 1949) makes it possible to deduce the information capacity of a communication system in which isolated pulse transmitters share a common channel without synchronization. Under these conditions it is possible to add redundant information at the output of each transmitter and to separate the signals at the receiver in much the same manner as a wanted signal is separated from random noise. Some simple systems are discussed and specimen numerical data for a 6-station multiplex system are tabulated. Applications of such a system, especially to mobile communication, are suggested.

621.396.5 : 621.396.619.16 1768

**Pulse Multiplex Equipment for 24-Channel Telephony.**—In the journal reference of 1908 of April please read Vol. 30 for Vol. 29.

621.396.619.16 1769

**Pulse Modulation.**—E. Prokott. (*Arch. elekt. Übertragung*, Jan. 1950, Vol. 4, No. 1, pp. 1–10.) Description of the basic methods: p.a.m., p.ph.m., p.f.m. and p.w.m. The latter is considered as higher-order amplitude modulation. The sideband theory of the different systems is developed, with particular reference to the structure of the pulse train and to application possibilities.

621.396.619.16 1770

**Interference Characteristics of Pulse-Time Modulation.**—E. R. Kretzmer. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 252–255.) 1949 I.R.E. National Convention paper. Mathematical and experimental analysis, with particular reference to pulse-duration and pulse-position modulation. Both 2-station and 2-path interference are considered. In 2-station interference the stronger signal is completely predominant and the noise has a random character. For single-channel pulse-duration modulation, 2-path interference generally permits fairly good reception of speech and music signals.

621.396.619.16 1771

**On Distortion in Pulse-Width Modulation.**—J. Müller. (*Arch. elekt. Übertragung*, Feb. 1950, Vol. 4, No. 2, pp. 51–58.) An investigation of the deformation of rectangular pulses due to the limited receiver bandwidth, as dependent on the ratio $\pi$ of pulse period to pulse duration, and also of the distortion caused by subsequent amplitude limiting. Fourier analysis is used and, in an alternative method, the transfer functions of Küpfmüller. The bandwidth necessary for p.w.m. is determined and the relation of bandwidth to pulse ratio $\pi$ is considered. Experiments confirm the theory.

621.396.65 1772

**Problems of Decimetre-Wave Directional Links.**—G. Megla. (*Elektrotechnik, Berlin*, April 1948, Vol. 2, No. 4, pp. 103–109.) Discussion of fundamental considerations in the technique, including choice of wavelength, aerial systems and bandwidth, height of stations, signal level, sensitivity, aerial gain.

621.396.712 : 621.397.7 1773

**WTCN A.M. F.M. TV.**—J. M. Sherman. (*Broadcast News*, Jan./Feb. 1950, No. 57, pp. 32–39.) Illustrated description of studio and transmitter facilities. The aerials and microwave relay receivers are installed at the top of the Foshay tower, the tallest building in the Minneapolis-St Paul area.

621.396.712 1774

**KENI, Anchorage, Alaska.**—A. G. Hiebert. (*Broadcast News*, Jan./Feb. 1950, No. 57, pp. 40–44.) Illustrated description of studio and transmitter equipment.

621.396.712 1775

**Radio Istanbul covers Near East with 150-kW Transmitter.**—P. C. Brown. (*Broadcast News*, Jan./Feb. 1950, No. 57, pp. 28–31.) Illustrations and a few details of equipment and studio facilities. The R.C.A. Type BTA-150A transmitter operates on 704 kc/s with a 725-ft aerial tower.

621.396.712 1776

**Voice of America's 150-kW Transmitter.**—E. L. Schacht. (*Broadcast News*, Jan./Feb. 1950, No. 57, pp. 24–27.) Brief details, with illustrations, of a new R.C.A. Type BTA-150A a.m. transmitter equipment installed in Munich, Germany, to relay on 1 195 kc/s the Voice-of-America programmes as they originate in the United States. The aerial system consists of a 4-element array of half-wave towers, giving three different beam patterns which can be selected by switching. The forward power gain on the main lobes is >6.

621.396.931 : 621.396.826 1777

**Echoes in Transmission at 450 Mc/s from Land- to Car-Radio Units.**—W. R. Young, Jr, & L. Y. Lacy. (*Proc. Inst. Radio Engrs*, March 1950, Vol. 38, No. 3, pp. 255–258.) 1949 I.R.E. National Convention paper. Examples of multipath echoes observed in a moving car in New York City are shown and discussed.

## SUBSIDIARY APPARATUS

621-526 : 621.3.016.35 1778

**Contribution to the Study of the Stability of Regulating Circuits and of Servomechanisms.**—P. L. Dubois-Violette. (*C. R. Acad. Sci., Paris*, 12th April 1950, Vol. 230, No. 15, pp. 1380–1383.)

621.396.682 : 621.316.722 1779

**Cathode Heater Compensation as Applied to Degenerative Voltage-Stabilized Direct-Current Power Supplies.**—R. C. Ellenwood & H. E. Sorrows. (*Bur. Stand. J. Res.*, Sept. 1949, Vol. 43, No. 3, pp. 251–255.) Full description of the method noted in 468 of February.

## TELEVISION AND PHOTOTELEGRAPHY

621.397.331.2 1780

**TV Camera-Tube Design.**—A. Lytel. (*TV Engng, N.Y.*, Jan.–March 1950, Vol. 1, Nos. 1–3, pp. 28–29, 22–25 & 24–26.) Mechanical, electronic and optical characteristics of nine types of pickup tube developed during the past twenty years.

621.397.6 : 621.385 1781

**Encyclopaedia of New and Available Components for Television Equipment.**—M. Alixant. (*Télévis. franç.*, April 1950, No. 57, pp. 9–12.) The first of a series of reviews. American and European c.r. tubes, receiving, rectifying, and transmitting valves are tabulated with their characteristics.

621.397.62 + 621.397.6.001.4 1782

**The Video-Frequency Stage.**—R. Gondry. (*Télévis. franç.*, Nov. & Dec. 1949, Nos. 53 & 54, pp. 6–9 & 13–17.) Possible circuit arrangements and adjustments for near-linear frequency response are discussed. Circuits of two useful test instruments are given: (*a*) a square-wave generator; (*b*) a tuned amplifier.

621.397.62 1783

**Characteristics of High-Efficiency Deflection and High-Voltage Supply Systems for Kinescopes.**—O. H. Schade. (*RCA Rev.*, March 1950, Vol. 11, No. 1, pp. 5–37.) The losses in the separate components of reactive deflection systems are examined individually and in relation to each other. By reducing valve and circuit losses, combined h.v. and deflection systems can be produced having high efficiency and low cost.

621.397.62 1784

**Simplified Intercarrier Sound.**—W. J. Stroh. (*Electronics*, April 1950, Vol. 23, No. 4, pp. 106–109.) By using the new Type-6BN6 valve (1292 of May) as the limiter detector of an intercarrier sound system, a considerable reduction can be effected in the number of valves and tuned circuits of a television receiver. Such equipment compares favourably with other f.m. detectors as regards suppression of ignition interference and of the a.m. which usually accompanies f.m.

621.397.62 1785

**TV Applications of the 6BN6.**—R. O. Gray & W. J. Stroh. (*FM-TV*, March 1950, Vol. 10, No. 3, pp. 14–15, 30.) When used as an audio limiter-discriminator this valve replaces 3 conventional types. As a synchronization separator it also economizes in components, space, and wiring time. See also 1292 of May (Adler).

621.397.62 1786

**A New Method of Synchronisation for Long-Range TV Reception.**—T. B. Tomlinson. (*Electronic Engng.*, April 1950, Vol. 22, No. 266, pp. 149–151.) The most objectionable feature of long-distance television reception is irregular triggering of the line timebase by noise in the synchronizing pulses. To overcome this a selective amplifier is used which extracts the fundamental component from the synchronizing pulse, which is then squared, differentiated and applied to the timebase. This system has been used outside the normal service area and consistently provides pictures with good entertainment value.

621.397.62 1787

**Permanent-Magnet Lenses for Television Tubes.**—D. Hadfield. (*Electronic Engng.*, April 1950, Vol. 22, No. 266, pp. 132–138.) Permanent-magnet focusing systems are becoming increasingly popular for economic reasons. Methods of investigating the field and electron-optical characteristics are described. Magnet design, choice of materials and methods of varying the field and of magnetizing are considered.

621.397.62 1788

**Technique and Developments of High-Definition Television Receivers.**—In the journal reference of 1018 of April please read Vol. 30 for Vol. 29.

621.397.62 1789

**A.V.C. System for Positive-Modulation Television.**—E. G. Beard. (*Philips tech. Commun., Aust.*, 1949, No. 6, pp. 9–14.)

621.397.62 1790

**Murphy V150 Television Set.**—(*Wireless World*, April 1950, Vol. 56, No. 4, pp. 133–136.) Test report of a table model with unusual features and using a 12-in. c.r. tube.

621.385.2 : 621.315.59 1791

**Use of Germanium Diodes at High Frequencies.**—J. H. Sweeney. (*Elect. Engng, N.Y.*, March 1950, Vol. 69, No. 3, pp. 217–220.) A.I.E.E. Winter Meeting paper, 1950. Ge diodes may be substituted for vacuum tubes in detector, d.c.-restorer and mixer circuits if due allowance is made for the different characteristics of the two types of diode. Compared with the vacuum diode, the Ge unit has a greater forward conductance, though it has a finite back resistance. It has less shunt capacitance and has zero current flow at zero voltage. Its resistance in either direction varies with temperature.

621.397.62 : 621.396.662 1792

**An Experimental Ultra-High-Frequency Television Tuner.**—T. Murakami. (*RCA Rev.*, March 1950, Vol. 11, No. 1, pp. 68–79.) Circuits and detailed description of an experimental 500–700-Mc/s tuner to enable field tests to be made by standard television receivers on television transmission from the N.B.C. u.h.f. station, Bridgeport (see 1795 below). Each tuning element consists of two strips of copper foil on a cylindrical former, tuned by means of a copper or brass core. A double-heterodyne system converts the u.h.f. signals to the 21–27-Mc/s intermediate frequency of the standard television receiver. Performance figures are given.

621.397.62 : 621.396.662 1793

**Continuous Tuning for TV Sets.**—French. (*See* 1762.)

621.397.7 + 621.396.712 1794

**WTCN A.M. F.M. TV.**—Sherman. (*See* 1773.)

621.397.7 1795

**Experimental Ultra-High-Frequency Television Station in the Bridgeport, Connecticut Area.**—R. F. Guy, J. L. Seibert & F. W. Smith. (*RCA Rev.*, March 1950, Vol. 11, No. 1, pp. 55–67.) The first of a series of reports on the N.B.C. transmitter KC2XAK. Visual and aural signals from station WNBT, distant 54 miles, are received by a parabolic aerial and retransmitted on carrier frequencies of 530.25 and 534.75 Mc/s respectively, using a pylon slot aerial 330 ft above average ground level. Each transmitter will give power outputs of 0.5 kW aural and 1.0 kW peak visual. With an aerial gain of 17, the effective radiated power of the system is approximately 13.9 kW peak visual, the overall efficiency being approximately 80%.

621.397.7 1796

**WHIO-TV, Dayton.**—(*Broadcast News*, Jan.–Feb. 1950, No. 57, pp. 45–57.) A fully illustrated description of the lay-out and equipment of the studios of the Miami Valley Broadcasting Corporation.

621.397.7 : 621.397.81 1797

**U.H.F. Television Field Test.**—R. F. Guy. (*Electronics*, April 1950, Vol. 23, No. 4, pp. 70–76.) A description of experimental equipment for obtaining data on range, field strength and interference effects in u.h.f. television reception. A 500-W commercial v.h.f. unit operating at 176.75 Mc/s is followed by a tripler and amplifier. A vestigial sideband filter is provided, picture and sound signals being combined in a notch filter and fed to the aerial through a single coaxial line. A unidirectional slot aerial gives a vertical beam-width of about 2° for the main lobe at half-power points, and a power gain > 17. Conventional valve technique is satisfactorily employed in the receiving equipment, which includes a special tuner and converter covering the range 500–700 Mc/s. See also 1792 and 1795 above.

621.397.828 1798

**A Study of Co-Channel and Adjacent-Channel Interference of Television Signals.**—R.C.A. Laboratories Division. (*RCA Rev.*, March 1950, Vol. 11, No. 1, pp. 99–120.) Television carrier synchronization has been found extremely advantageous in reducing co-channel interference. Offset-carrier operation gives even better results and is more economical to apply. Moreover, it is equally applicable to standard monochrome transmissions and to the dot-sequential and the field-sequential colour systems. For either monochrome or colour transmissions it is recommended that the carrier offset shall be 10.5 kc/s, in a 3-station combination one being above and another below the assigned frequency.

621.397.6 1799

**Television for Radiomen.** [Book Review]—E. M. Noll. Publishers: Macmillan & Co., London, 595 pp., 52s. 6d. (*Wireless Engr*, April 1950, Vol. 27, No. 319, p. 134.) "It covers all aspects of television receivers in an elementary manner and includes transmitting material only in just enough detail to give the reader a picture of the complete system . . . The book should prove extremely valuable to those for whom it is intended, that is, those who have to maintain and repair American television receivers. There is a good deal of material in it useful to those who handle British receivers."

## VALVES AND THERMIONICS

621.383.032.217 1800

**Optical and Photoelectric Properties of Antimony-Caesium Cathodes.**—H. D. Morgulis, P. G. Borzyak & B. I. Dyatlovitskaya. (*C. R. Acad. Sci. U.R.S.S.*, 21st June 1947, Vol. 56, No. 9, pp. 925–928.)

621.385 1801

**Ribbon-Beam Valves.**—J. L. H. Jonker. (*Tijdschr. ned. Radiogenoot.*, March 1950, Vol. 15, No. 2, pp. 37–52. In Dutch, with English summary.) Results are presented of laboratory work on the development of special-purpose beam-deflection valves using ribbon beams, particularly for rapid switching, as in telephony. The replacement of the more common pencil beam by the ribbon beam permits designs having dimensions and anode voltages comparable with those of ordinary receiving valves.

621.385 1802

**Miniature Valves in the French Radio Industry.**—L. Thibierge & H. Saucet. (*Tech. mod., Paris*, 1st/15th March 1950, Vol. 42, Nos. 5/6, pp. 65–73.) Illustrated account of the development, manufacture and testing of 7-pin miniature valves in the works of the Société Française Radio-Électrique.

621.385 : 621.396.621 1803

**New One-Tube Limiter-Discriminator for F.M.**—A. P. Hasse. (*Tele-Tech*, Jan. & Feb. 1950, Vol. 9, Nos. 1 & 2, pp. 21–23, 49 & 32–33, 61.) Construction, characteristics and applications of the Type-6BN6 gated-beam valve. For another account see 1292 of May (Adler).

621.385 (083.72) 1804

**Standards on Electron Tubes: Definitions of Terms, 1950.**—(*Proc. Inst. Radio Engrs*, April 1950, Vol. 38, No. 4, pp. 426–438.) A comprehensive list prepared by a committee of the Institute of Radio Engineers. Definitions of valve parameters have been generalized to apply to all valve types at any frequency, and definitions pertaining to beam valves are included.

621.385.012 : 621.3.032.213.2 1805

**The Influence of the Heating on Valve Cathode Current.**—G. Kessler. (*Arch. Elektrotech.*, 1950, Vol. 39, No. 9, pp. 601–619.) The relation between filament power and emitter temperature is established, and the effect of temperature change on cathode current is determined and expressed as an equivalent grid-voltage variation. The derived relations are confirmed by measurements, evaluated numerically, and applied to discussion of the design of amplifiers of improved performance at very low frequencies, e.g. $<10$ c/s.

621.385.015.3 1806

**Some Calculations of Transients in an Electronic Valve.**—D. R. Hartree. (*Appl. sci. Res.*, 1950, Vol. B1, No. 5, pp. 379–390.) "The integro-differential equation for the non-steady behaviour of a beam of electrons injected into the space between two plane parallel electrodes at given potentials is obtained. When the anode and grid are kept at the same potential $V_a$, the leading electrons reach the anode with energies greater than $eV_a$. If the front of the beam is sharp and the injected beam current is constant, then for the period until some electrons reach the anode and while the spatial order of electrons in the inter-electrode space remains the same as the time-order in which they passed through the grid, the integro-differential equation has an exact solution in terms of Airy functions. An example is given of results calculated from this solution, where applicable, and continued by numerical integration of the integro-differential equation."

621.385.029.6 : 621.396.615.14 1807

**On the Physics and Technics of Modern Transmitting Valves for Ultra High Frequencies.**—K. W. Reusse. (*Elektrotechnik, Berlin*, Feb. & March 1950, Vol. 4, Nos. 2 & 3, pp. 33–42 & 81–90.)

621.385.029.63/.64 : 621.396.615.14 1808

**The Amplification of Centimetre Waves: Travelling-Wave Valves.**—In the journal reference of 1294 of May please read Vol. 30 for Vol. 29.

621.385.029.63/.64 1809

**Study of the Different Waves that can be propagated along a Line in Interaction with an Electron Beam. Application to the Theory of the Travelling-Wave Amplifier.**—P. Lapostolle. (*Ann. Télécommun.*, Feb. & March 1948, Vol. 3, Nos. 2 & 3, pp. 57–71 & 85–104.) Detailed discussion of the possible waves in such systems and, in particular, of the wave whose amplitude increases along the line for certain values of the beam velocity. The study of particular models is generalized to apply to any type of delay line. The theory is applicable to beams of any diameter, intensity and velocity, and line losses are taken into account. Curves are provided which facilitate the solution of many problems associated with the specially important case of very intense beams.

621.385.029.63/.64 1810

**Traveling-Wave Tubes.**—J. R. Pierce. (*Bell Syst. tech. J.*, Jan. 1950, Vol. 29, No. 1, pp. 1–59.) Part 1 of a four-part presentation of the contents of a forthcoming book. Chapter 1 is a brief introduction which stresses the great practical advantage of very large bandwidth offered by the travelling-wave amplifier. Chapter 2 presents a simple theory of the gain for small signals on the assumption that all the electrons are acted on by the same a.c. field and are displaced by the field in the axial direction only; the gain is derived from the propagation constant of the forward wave with increasing amplitude. Chapter 3 discusses the helix, a high-impedance circuit capable of propagating slow waves with a phase velocity which is substantially constant over a wide frequency range. An approximation to the properties of a helix is obtained by considering a helically conducting cylindrical sheet of the same radius and pitch. Curves are given which enable the propagation constant, wave velocity, gain parameter and impedance parameter to be determined. The properties of a wire helix are discussed by 'developing' the helix into a flat sheet and considering the special cases of two turns or four turns per wavelength. Optimum design of the helix is treated in terms of the ratios of the beam radius to helix radius and of the wire diameter to pitch. The chapter concludes with a consideration of the transmission-line equivalent to a helix and of losses. Appendix 1 summarizes the field equations and the properties of the Bessel functions I and K and gives curves showing various series approximations to $I_0$, $I_1$, $K_0$ and $K_1$. Appendix 2 discusses propagation on a helically conducting cylinder.

621.385.029.64/.65 1811

**Helix Parameters Used in Traveling-Wave-Tube Theory.**—R. C. Fletcher. (*Proc. Inst. Radio Engrs*, April 1950, Vol. 38, No. 4, pp. 413–417.) Helix parameters used in the normal-mode solution for the travelling-wave valve are determined by considering the field equations for a thin electron beam, the two methods of treatment being found equivalent. Corresponding parameters for a thick electron beam are found by considering an equivalent thin beam with approximately the same r.f. admittance.

621.385.032.213.2 : 621.317.336.1 1812

**Change of Mutual Conductance with Frequency.**—A. Eisenstein. (*Wireless Engr*, March 1950, Vol. 27, No. 318, pp. 100–101.) Comment on 1301 of May (Raudorf). Experiments in which square voltage pulses were applied to the grid of a pentode are discussed. The results obtained support the view that an increase of cathode-interface thickness with valve life offers an explanation both of the observed frequency dependence of the mutual conductance $g_m$ and of the reduction in capacitance with increasing resistance reported by Raudorf. See also 490 of February.

621.385.032.216 1813

**Vacuum Factor of the Oxide-Cathode Valve.**—G. H. Metson. (*Brit. J. appl. Phys.*, March 1950, Vol. 1, No. 3, pp. 73–77.) Measurements on a variety of new valves showed wide variations of vacuum factor $k$, but these values fell after a period of operation to an approximately constant value $k_0$ in the range 300–900 $\mu\mu$A/mA. The variations of $k_0$ were examined with respect to anode voltage, anode current, and to cathode and envelope temperatures. A theory is proposed to explain the anomalous form of variation with anode voltage, which has a bearing on the interpretation of ionization-gauge measurements for pressures $< 10^{-6}$ mm Hg.

621.385.15 1814

**Receiving Tubes employing Secondary-Electron Emitting Surfaces exposed to the Evaporation from Oxide Cathodes.**—C. W. Mueller. (*Proc. Inst. Radio Engrs*, Feb. 1950, Vol. 38, No. 2, pp. 159–164.) The poisoning of the secondary emitting surface due to cathode evaporation is reduced by careful control of cathode materials and temperature, so that the surface can be exposed directly to the cathode evaporation. This evaporation, if not excessive, may even enhance the secondary emission. The construction and characteristics are described of a valve with a transconductance of 24 mA/V, a wide-band figure-of-merit two to three times that of conventional valves such as Types 6AG5 and 6AK5, and a life of at least 2 000 hours.

621.385.15.001.8 1815

**Recent Applications of Electron Multiplier Tubes.**—J. S. Allen. (*Proc. Inst. Radio Engrs*, April 1950, Vol. 38, No. 4, pp. 346–358.) Applications to the measurement of very small currents or the counting of single ions or electrons are considered. The characteristics of secondary-electron emitting surfaces suitable for use in multiplier tubes are discussed and recently developed tubes with one or more stages of electron multiplication are described. The statistical treatment of the size distribution of pulses from multiplier tubes is also discussed.

621.385.2 1816

**The Current Build-Up in a Planar Diode.**—E. H. Gamble. (*J. appl. Phys.*, Feb. 1950, Vol. 21, No. 2, pp. 108–112.) An analytical study of the development of the space-charge cloud of electrons and its motion across the interelectrode space, and of the growth of current in the external circuit. Under space-charge-limited conditions the initial build-up is a function of higher powers of time than under temperature-limited conditions. The build-up may be considerably delayed or hastened by application of sine-wave signals of suitable amplitude.

621.385.2/.3 : 621.315.59 1817

**Observations of the Rapid Withdrawal of Stored Holes from Germanium Transistors and Varistors.**—L. A. Meacham & S. E. Michaels. (*Phys. Rev.*, 15th April 1950, Vol. 78, No. 2, pp. 175–176.) Investigations of transistor response to square pulses of applied voltage have shown bursts of collector current provoked by preceding emitter current in units employing N-type germanium. These are thought to be due to the rapid withdrawal of stored holes produced by the forward emitter current and existing in the germanium at the time of application of the collector potential.

621.385.2 : 621.315.59 1818

**On the Surface Conductance of Germanium.**—P. Aigrain. (*C. R. Acad. Sci., Paris*, 20th Feb. 1950, Vol. 230, No. 8, pp. 732–733.) The surface conductance of a diode of 20-kΩ back resistance was demonstrated by measuring voltage variations at two beryllium-bronze contact points equidistant from the input point but on opposite faces of the Ge disk; except at very low input currents the variation on the input face was much the greater. Since this variation depends only on input current, the transistor effect must be due to a modulation of the surface conductance. See also 1305 of May.

621.385.2 : 621.315.59] : 621.397.62 1819

**Use of Germanium Diodes at High Frequencies.**—Sweeney. (*See* 1791.)

621.385.3 : 621.396.615.14 1820

**Electron Transit Time.**—S. K. Chatterjee & B. V. Sreekantan. (*Wireless Engr*, Feb. 1950, Vol. 27, No. 317, pp. 59–63.) Transit-time effects on the efficiency and limiting oscillation frequency were studied for Type 833A, 834, 304B, 316A, and 955 valves, using Gavin's expression for transit time (4358 of 1939), modified to take account of the negative grid bias and the alternating voltages on the anode and grid. The limiting frequency is reached for the first three of the above valves when the oscillation period is nearly four times the electron transit time from cathode to anode, and for the other two valves when the period is nearly three times the transit time.

621.385.4 1821

**Aligned-Grid Valves.**—D. C. Rogers. (*Wireless Engr*, Feb. 1950, Vol. 27, No. 317, pp. 39–46.) An experimental investigation was made of the distribution of current density in the plane of the screen grid. In order to check the validity of existing assumptions about aligned-grid valves, and to determine the optimum distance between the control and screen grids, a demountable planar valve was used in which the screen grid was replaced by a plate with a 0.1-mm slot parallel to the grid wires. The current flowing to a collector electrode behind the slot was measured. Electrode distances were adjustable by means of micrometer screws and the slotted plate had similar adjustment in its plane at right angles to the slot. The control-grid pitch and total anode current were kept constant and the collector current was measured across the beam from the centre grid aperture at intervals of 0.1 mm, the grid being at cathode potential. The process was repeated for various electrode distances. The results indicate that for minimum screen current the screen should be as close as possible to the control grid and not, as is often assumed, in the plane of focus of the electron beam.

621.385.83 : 621.396.619.23 1822

**The Electron Coupler: a Spiral-Beam U.H.F. Modulator.**—C. L. Cuccia & J. S. Donal, Jr. (*Electronics*, March 1950, Vol. 23, No. 3, pp. 80–85.) A description based on that given in a 1949 I.R.E. National Convention paper. See 2975 of 1949.

621.385.832 1823

**A Method of Reducing Deflection Defocusing in Cathode-Ray Tubes.**—R. G. E. Hutter & S. W. Harrison. (*J. appl. Phys.*, Feb. 1950, Vol. 21, No. 2, pp. 84–89.) The causes of spot distortion are explained and an electron gun is described in which beam predistortion is used for spot correction. The method can be used for both e.s. and e.m. focusing and deflecting systems.

621.396.615.14 1824

**Observations on Some Electronic Fundamentals of Microwave Valves.**—W. Sigrist. (*Bull. schweiz. elektrotech. Ver.*, 21st Jan. 1950, Vol. 41, No. 2, pp. 35–42. In German.) A graphical method of integration is described by which the electron paths may be exactly traced after the fields at the electrodes have been determined by the use of an electrolyte tank.

1825

621.396.615.14

**The Linear Theory of Transit-Time Valves.**—F. W. Gundlach. (*Fernmeldetech. Z.*, Oct. 1949, Vol. 2, No. 10, pp. 319–328.) The operation of almost all transit-time valves can be referred to the fundamental problem of electron motion in a plane capacitor. Under the assumption of small alternating voltages and currents, analytical relations are derived for the electron motion and for the currents thereby produced in the capacitor. The results are applied to the diode oscillator, the retarding-field valve, the klystron and its reflex type, and to the Heil generator.

1826

621.396.615.141.4

**The Flow of Standing Current in a Cylindrical Whole-Anode Magnetron.**—V. M. Glagolev. (*Zh. tekh. Fiz.*, Aug. 1949, Vol. 19, No. 8, pp. 943–951. In Russian.)

1827

621.396.615.141.2

**Phase Focusing in a Magnetron with Plane Electrodes.**—B. M. Zamorozkov. (*Zh. tekh. Fiz.*, Nov. 1949, Vol. 19, No. 11, pp. 1321–1328. In Russian.) The idea of phase focusing was used by Gobubkov in 1939 (*Bull. Acad. Sci. U.R.S.S.*, 1940, Vol. 4, p. 524) for providing a general theory which would be applicable to various types of u.h.f. oscillators. This theory is applied to a study of the magnetron. It is fully adequate for interpreting the operation of the magnetron and equation (11) is derived determining the critical conditions. The relation between the generated wavelength and the intensity of the magnetic field is determined from the general equation (13) of Okabe (*Radio Res. Rep. Japan*, 1935, Vol. 5, p. 69). The possibility is also suggested of the appearance of second-order foci and of oscillations due to interaction between these foci and the electrodes. The theoretical results have been confirmed experimentally.

1828

621.396.615.141.2

**The Planar Magnetron under Static Conditions of Space Charge.**—J. L. Delcroix & G. A. Boutry. (*C. R. Acad. Sci., Paris*, 13th March 1950, Vol. 230, No. 11, pp. 1046–1048.) The equations are stated for space-charge conditions when the anode potential is higher than the cut-off potential. When cut-off occurs, the space-charge conditions can be deduced from those for the preceding case. The electron paths in the two cases are arcs of one and the same geometric curve.

1829

621.396.615.141.2 : 621.396.643

**A New Type of Magnetron Amplifier.**—P. Marié. (*Onde élect.*, Jan., Feb. & April 1950, Vol. 30, Nos. 274, 275 & 277, pp. 13–22, 79–90 & 200–202.) The variation of a negative resistance as a function of voltage is studied, the internal mechanism of the magnetron is discussed, the threshold resistance calculated and the relative amplitudes of the reflected and incident waves are determined. The threshold resistance decreases with frequency and in a wide band centred on the mean frequency can be considered as the reciprocal of the characteristic impedance of a waveguide with a given cut-off frequency. Separation of the reflected and incident waves requires the generation of polyphase waves. Their theory and the construction of 3-phase directional couplers are discussed. The constructional details of the magnetron itself are described. It is of the split-anode type, with 12 anode segments strapped in interleaved groups of 4. Using this system of overlapping anodes, the directional coupling may be incorporated in the valve itself. Experimental results will be given later.

1830

621.396.615.142.2

**On the Theory of the Klystron.**—V. N. Shevchik. (*Zh. tekh. Fiz.*, Nov. 1949, Vol. 19, No. 11, pp. 1271–1275. In Russian.)

1831

621.396.822

**Spontaneous Fluctuations in Double-Cathode Valves.**—R. Fürth. (*Wireless Engr*, April 1950, Vol. 27, No. 319, p. 129.) Corrections to paper abstracted in 1033 of April.

1832

621.396.822

**Are Transit-Angle Functions Fourier Transforms?**—W. E. Benham. (*Wireless Engr*, April 1950, Vol. 27, No. 319, pp. 131–132.) Further discussion. See 2981 of 1949.

1833

621.396.822

**Induced Grid Noise.**—R. L. Bell. (*Wireless Engr*, March 1950, Vol. 27, No. 318, pp. 86–94.) Under favourable conditions there is an exact relationship between mean-square noise currents induced at the grid of a triode valve and the space-charge component of input capacitance measurable at that point. In practice this provides an approximate relation useful for estimating the total mean-square induced noise from other measurements. Techniques of grid-noise measurement are discussed and attention is drawn to the occurrence of various inherent errors. Results are presented of measurements of induced grid noise in microwave triodes at 200 Mc/s. Available three-halves power-law treatments are inadequate to explain the induction of shot noise at a control grid, but values of space-charge admittance measured on the valve together with other low-frequency measurements may be made to yield much closer estimates of induced grid noise in any structure, for moderate transit angles.

1834

621.396.822

**Fluctuation Process as Oscillations with Random Amplitude and Phase.**—V. I. Bunimovich. (*Zh. tekh. Fiz.*, Nov. 1949, Vol. 19, No. 11, pp. 1231–1259. In Russian.) It is known that fluctuations at the output of an oscillatory system have the form of sinusoidal oscillations with amplitude and phase varying more or less slowly. In many cases it is convenient to represent fluctuations in the form $E\cos\phi$ where $E$ and $\phi$ have definite and statistically independent random values. An investigation is presented of the statistical properties of the amplitude and phase of a fluctuation process. The investigation is divided into the following sections: (*a*) envelope; (*b*) fundamental relations; envelope and phase of a random process; (*c*) laws governing the probability distribution; (*d*) probability distribution for the phase of fluctuations.

1835

621.385

**Rundfunkröhren, Eigenschaften und Anwendung (Properties and Applications of Broadcasting Valves).** [Book Review]—L. Ratheiser, revised and enlarged by H. Hönger & G. Hinke. Publisher: Regelieus Verlag, Berlin-Grunewald, 1949, 440 pp., 27DM. (*Arch. elekt. Übertragung*, Jan. 1950, Vol. 4, No. 1, p. 32.) "Helpful for all concerned with broadcast reception."

## MISCELLANEOUS

1836

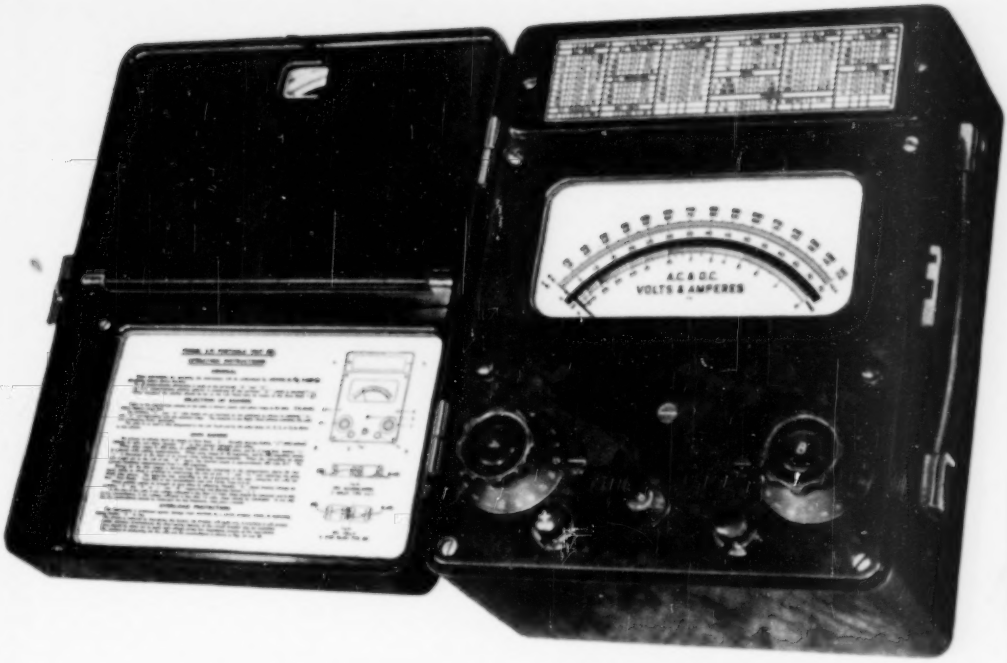
621.396.6(083.75)

**Standards on Designations for Electrical, Electronic and Mechanical Parts and their Symbols, 1949.**—(*Proc. Inst. Radio Engrs*, Feb. 1950, Vol. 38, No. 2, pp. 118–124.)

1837

621.396.6.001.2 : 621.396.9 : 629.13.052

**Mechanical Development of a Radio Altimeter.**—R. T. Croft. (*J. Brit. Instn Radio Engrs*, Feb. 1950, Vol. 10, No. 2, pp. 62–71.) Discussion of the problems involved in converting a laboratory model into an engineered instrument suitable for production in quantity. The general principles are illustrated by considering the detailed construction of a radio altimeter designed by the research staff of the Telecommunications Research Establishment.




# The WESTON S.75 Multi-Range Test Set

## 53 Ranges with Rotary Switch Selection

This uniquely comprehensive Test Set has 53 ranges for measuring AC and DC current and voltage, resistance and insulation. It is completely self-contained, with internal batteries to provide power for the ohms ranges and self-contained power pack for insulation measurement at 500 v. Selection is carried out by two 20-position switches. A fully protective safety device is fitted and is operative for forward or reverse overload. The 150 division 6" scale is uniformly divided and is fitted with an anti-parallax mirror. The set is enclosed in a handsome bakelite case and fully complies with B.S.S. No. 89 covering first grade instruments. Full details of this, and other Weston electrical measuring instruments will gladly be supplied on request.

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Type TF 912

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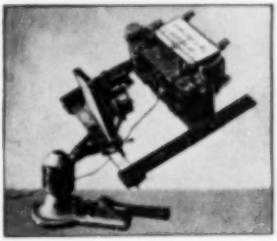

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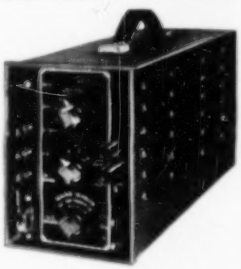

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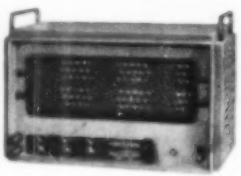

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*Type. O.222 7 Kc/s - 8 Mc/s*

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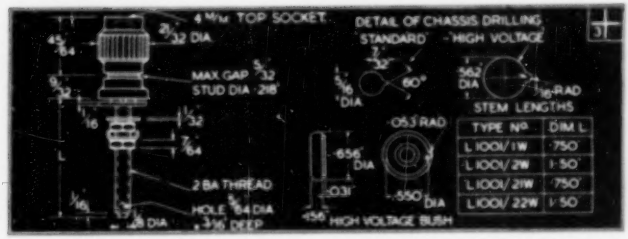
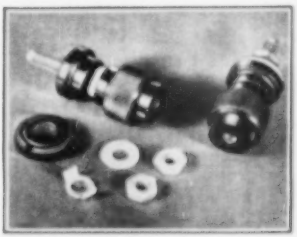
# The "Belling-Lee" page for Engineers

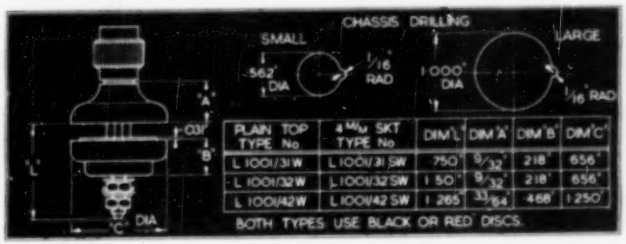


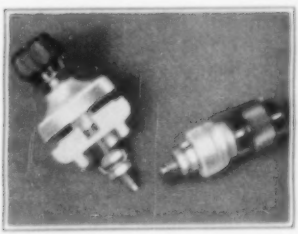



| Type | Rtg. in amperes | Peak Wkg. Voltage |
|---|---|---|
| "B" Standard | 15 | 1,000 |
| "B" High voltage | 15 | 2,000 |
| "H" | 100 | 1,500 |
| "F" | 30 | 1,000 |
| "L" Standard | 15 | 1,000 |
| "L" High voltage | 15 | 2,000 |
| "R" when fitted with insulating bush & washer | 9 | 350 |
| "W" | 15 | 750 |

The famous "Belling-Lee" "B" type terminal is now offered as a *High leakage resistance* terminal (details below). The above, the standard and high voltage types, can be supplied with plain red or black high grade bakelite tops* or with 4 mm. top socket, and have an anti-twist wedge on collar; no thread in clamping gap; special eared clamping washer and soldering spill for alternative methods of connection; nickel plated metal parts.

Ratings and voltages of these and other terminals are shown in the table opposite.



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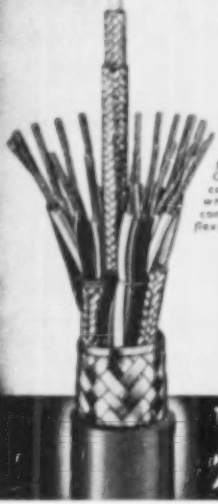

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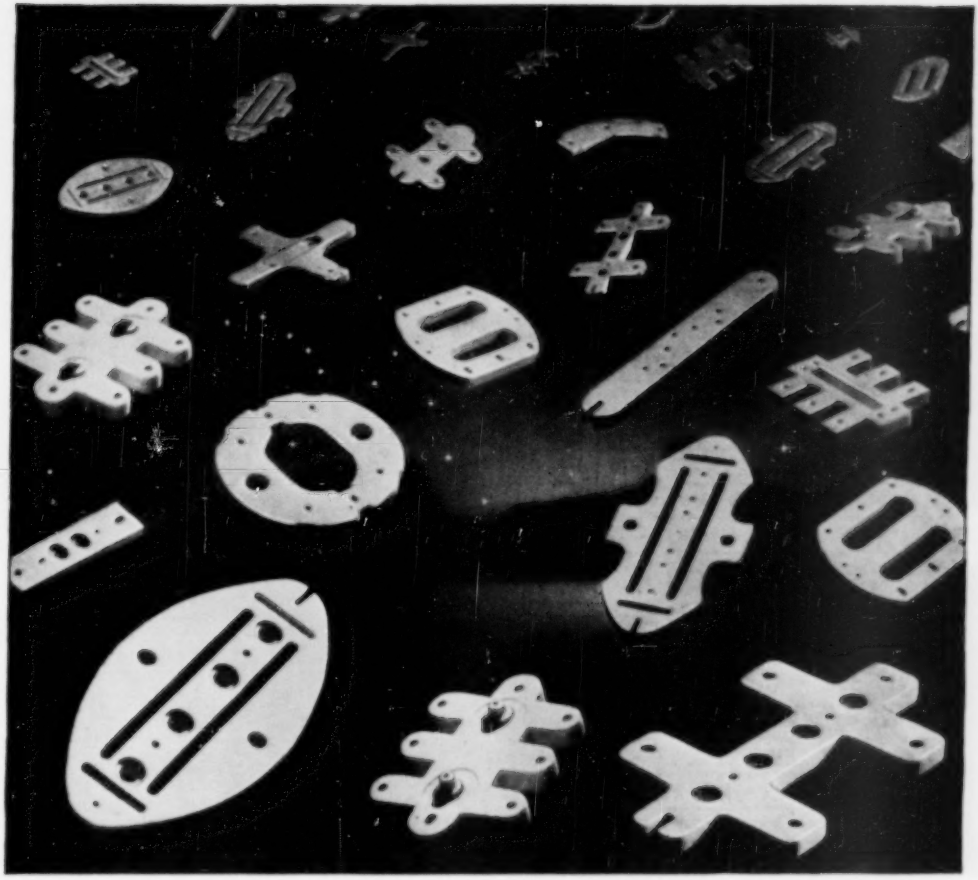

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| B3KO | 3" (sq.) | .57" | 7,500 | 14,000 | .5 w. |
| B4KO | 4" (sq.) | .57" | 7,500 | 14,000 | .75 w. |
| B5K | 5" | .57" | 7,500 | 14,000 | 2 w. |
| K5Q | 5" | .75" | 8,500 | 21,000 | 2.5 w. |
| K5G | 5" | .75" | 10,000 | 25,000 | 3 w. |
| K6D | 6.5" | .75" | 6,500 | 16,500 | 2.75 w. |
| K6Q | 6.5" | .75" | 8,500 | 22,000 | 3 w. |
| K6G | 6.5" | .75" | 10,000 | 26,000 | 3.5 w. |
| K46D | 6.25 x 4.25 (Oval) | .75" | 6,500 | 16,500 | 2.5 w. |
| K46Q | 6.25 x 4.25 (Oval) | .75" | 8,500 | 22,000 | 3 w. |
| K4 G | 6.25 x 4.25 (Oval) | .75" | 10,000 | 26,000 | 3 w. |
| F6JO | 6.5" | .6" | 9,000 | 18,500 | 3 w. |
| Z8D | 8" | 1" | 6,500 | 25,000 | 4 w. |
| Z8M | 8" | 1" | 8,000 | 31,000 | 4.5 w. |
| Z8R | 8" | 1" | 9,500 | 37,000 | 5 w. |
| Z10D | 10" | 1" | 6,500 | 25,000 | 6 w. |
| Z10M | 10" | 1" | 8,000 | 31,000 | 6.5 w. |
| Z10R | 10" | 1" | 9,500 | 37,000 | 7 w. |
| P44 | 12.2" | 1.5" | 10,000 | 60,000 | 10 w. |
| G.12 | 12.2" | 1.75" | 13,000 | 146,000 | 12 w. |

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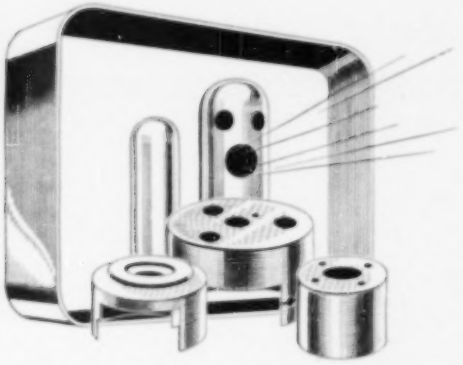

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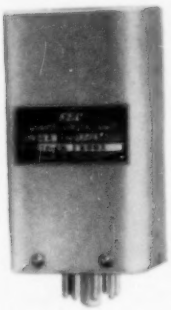

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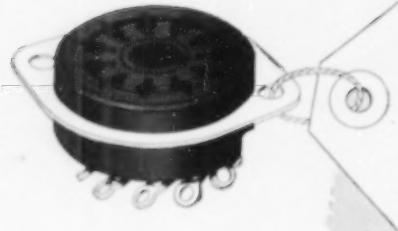

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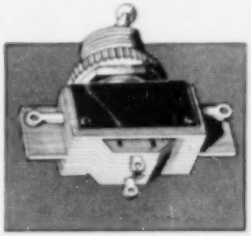

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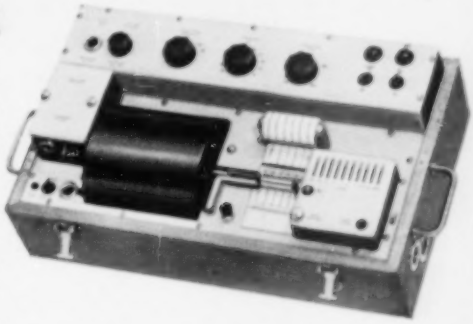

### *Specification*

**Frequency Range** : 20–150000 c/s.

**Zero Level** : 10 mV.

**Recording Speed** : Variable from 1000 to 50 db/sec with the 50-db potentiometer.

**Potentiometers** with 10, 25, 50 and 75 db ranges, a linear potentiometer from 10 to 110 mV and a phon potentiometer from 35 to 110 phons are obtainable.

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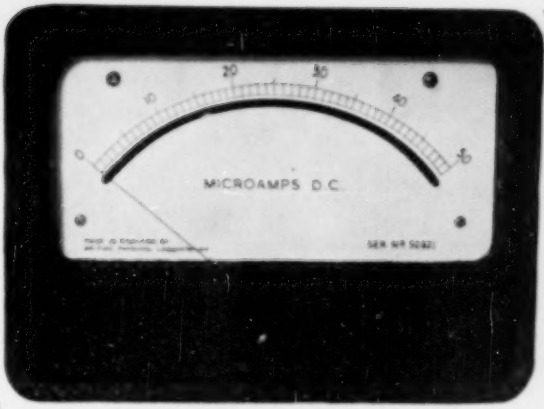


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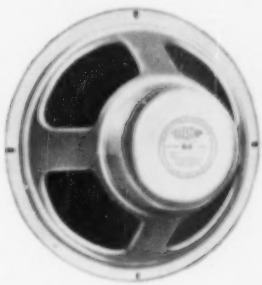

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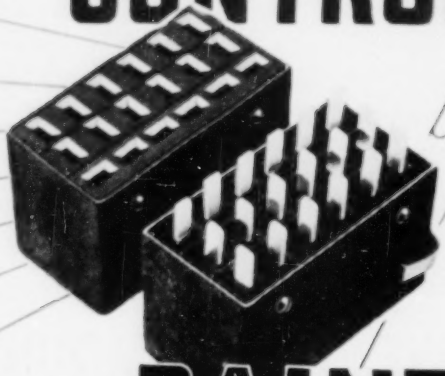

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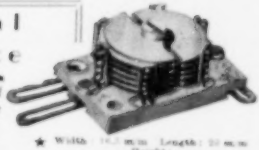

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## Index to Advertisers


| | PAGE |
|---|---|
| Acheson Colloids, Ltd. | 4 |
| Airmec Laboratories, Ltd. | 4 |
| All Power Transformer, Ltd. | 10 |
| Appointments | 26, 28 |
| Automatic Coil Winder & Elec. Equipt. Co., Ltd. | 8 |
| Belling & Lee, Ltd. | 21 |
| Birmingham Sound Reproducers, Ltd. | 6 |
| British N.S.F. Co., Ltd. | 3 |
| British Physical Laboratories | 24 |
| British Relay | 24 |
| Brookes Crystals, Ltd. | 30 |
| Bruel & Kjaer | 32 |
| Bulgin, A. F., Ltd. | Cover iii |
| Celestion, Ltd. | 29 |
| Cosmocord, Ltd. | 4 |
| De La Rue, Thomas, & Co., Ltd., Plastics Division | 7 |
| Dubilier Condenser Co. (1925) Ltd. | 1 |
| Edison Swan Electric Co., Ltd., The | 19 |
| Foyle, W., & G., Ltd. | 23 |
| Goodmans Industries, Ltd. | 16 |
| Griffiths, Gilbart, Lloyd, & Co., Ltd. | 24 |
| Guest, Keen & Nettlefolds, Ltd. | 11 |
| Harbors' Rubber Co., Ltd. | 30 |
| Hivac, Ltd. | 32 |
| Hunt, A. H., Ltd. | Cover iv |
| Lyons, Claude, Ltd. | 1, 27 |
| Marconi Instruments, Ltd. | 15 |
| McGraw-Hill Publishing Co., Ltd. | 25 |
| McMurdo Instrument Co., Ltd. | 30 |
| Muirhead & Co., Ltd. | 13 |
| Mullard Electronic Products, Ltd. | 12 |
| Muirex, Ltd. | 4 |
| Nagard, Ltd. | 22 |
| Oxley Development Co., Ltd. | 30 |
| Painton & Co., Ltd. | Cover ii |
| Pitman, Sir Isaac, & Sons, Ltd. | 24 |
| Products & Commerce, Ltd. | 24 |
| Reproducers & Amplifiers, Ltd. | 10 |
| Rothermel Electrical Products | 28 |
| Salford Electrical Instruments, Ltd. | 25 |
| Sangamo Weston, Ltd. | 2 |
| Spencer, Franklin, Ltd. | 24 |
| Steatite & Porcelain Products, Ltd. | 14 |
| Sullivan, H. W., Ltd. | 18 |
| Telegraph Condenser Co., Ltd., The | Cover iii |
| Telegraph Construction & Maintenance Co., Ltd., The | 22 |
| Telephone Mfg. Co., Ltd. | 5 |
| Vortexion, Ltd. | 18 |
| Wayne Kerr Laboratories, Ltd., The | 20 |
| Wego Condenser Co., Ltd. | 18 |
| Wimbledon Engineering Co., Ltd. | 6 |



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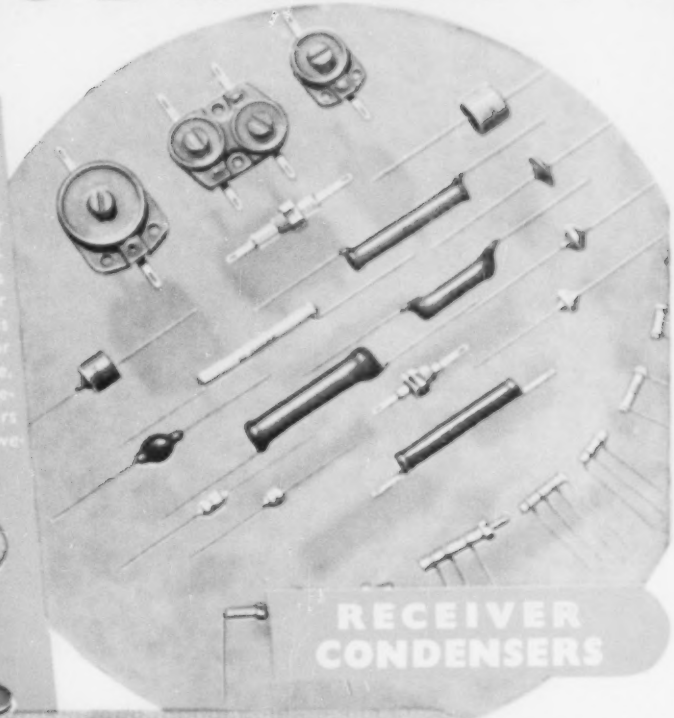

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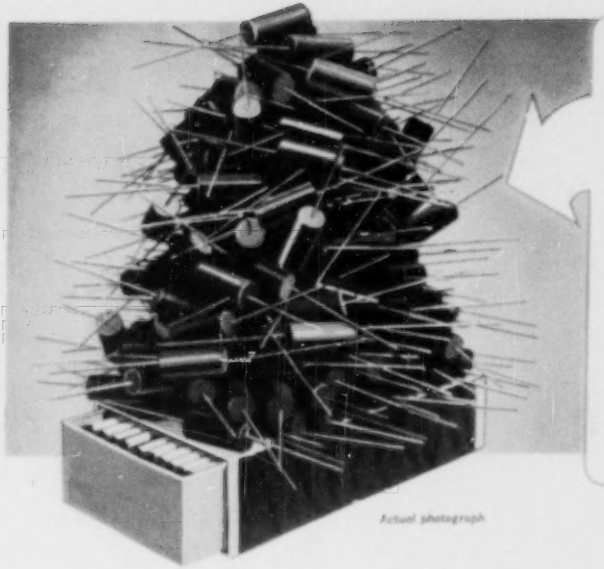

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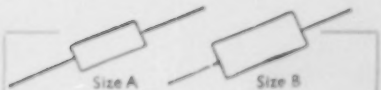

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